

**PHYSICAL MODELING OF THE IGBT AND APPLICATION FOR SMART
POWER IC DESIGN**

By

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Abstract of Dissertation Presented to the Graduate School
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This dissertation presents methodology for physical charge-based modeling of a lateral insulated-gate-bipolar transistor (LIGBT) for integrated circuit computer design. The steady-state and transient carrier transport problems in the LIGBT are formulated and solved. The effects of the two-dimensional current analysis in the lateral PNP BJT/LIGBT subcell is characterized, and the current-induced-space-charge region in the LDMOST/LIGBT subcell is modeled. These effects, which are not presented in conventional equivalent-(sub)circuit models, are physically and sometimes semi-numerically accounted for in our model. Improvements on both subcell models of the LIGBT have resulted in accurate simulation results without parameter optimization. The Kirk effect is modeled and added in our LIGBT model. Two-dimensional numerical device simulations were used extensively to study the effects and to aid the model

development. The LIGBT model is implemented in the Saber circuit simulator, by which the model equations are solved semi-numerically within the nodal analysis framework of the Saber. The parameter extraction method for Saber simulation is developed. Our new model is verified by measurements of the test devices, using simulations with model parameters extracted from static and dynamic measurements without any optimization to fit the measured data. The LIGBT model in Saber provides the capability for mixed-mode device/circuit simulation and hence can facilitate computer-aided optimal device/circuit design of IGBTs.

CHAPTER 1 INTRODUCTION

Recently, advanced process technologies of high-voltage (HV) devices have enabled the integration of HV devices with low-voltage control circuitry on the same chip. These technologies usually mix conventional bipolar and MOS processes and utilize advanced VLSI technology to create a very promising area in the IC industry. The application areas include high-voltage displays, motor controllers, regulators, TV circuits, telecommunication, switches, and interface-chips [BEC85, RUM85, ADL84].

The lateral insulated-gate bipolar transistor (LIGBT) is one of the MOS-controlled HV devices for power IC applications. The MOS channel in the LIGBT is usually formed by double-diffusion process; sequential diffusion of p- and n-type impurities in the epi region yields the MOS channel. The double-diffusion process is widely used for HV devices because of easier process steps to achieve short channels while realizing high breakdown voltages.

The LIGBT combines the advantages of current-carrying capability and the conductivity modulation of the IGBT with high input impedance and the high breakdown voltage of RESURF LDMOST. The on-resistance is about an order of magnitude lower than the LDMOST transistor of equivalent dimensions because the base

region is conductivity modulated. The cross-sectional view of the LIGBT device is shown in Fig.1.1. This device was fabricated with dielectrically isolated bipolar-CMOS-DMOS (BCDMOS) integrated-circuit technology developed at AT&T Bell Laboratories for HV applications [LU88]. Dielectric isolation (DI) effectively minimize parasitic cross coupling between circuit elements and to reduce the separation space between HV components compared to that which can be achieved with conventional junction isolation techniques.

Along with the matured process technology of the LIGBT and integrated circuits, the accurate CAD simulation tool for the LIGBT should be developed to support optimal device and circuit design. The circuit simulation for the LIGBT still relies on SPICE-based device models due to the unavailability of accurate the LIGBT device model [HEF93]. The SPICE-based models are unable to predict large overshoot voltages and currents and other switching effects critical in the design of power converters. Therefore, the development of physics-based model for the LIGBT and implementation into the circuit simulation tool, e.g., SABER, are needed. The advantages of a physics-based model are the accuracy of simulation results and the fundamental parameters which are available from layout and process information.

The purpose of this dissertation is to develop and implement a physical model for the LIGBT. The effects of the current-induced-space-charge region for the LDMOST/LIGBT subcell and two-dimensional current flow for a lateral PNP BJT/LIGBT subcell are modeled,

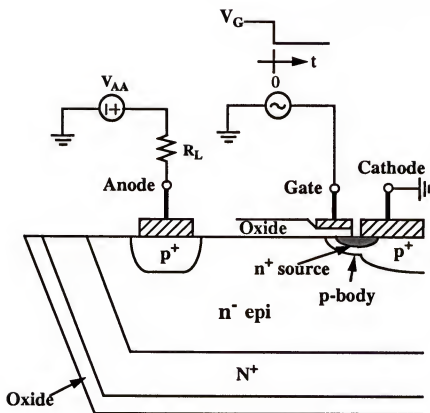


Fig.1.1 The cross sectional view of unit-cell of n-channel LIGBT. The series resistor load and gate drive test circuitry correspond to a transient turn-off from steady-on-state.

and the Kirk effect is characterized. The model parameter extraction method is developed and the physics-based LIGBT model is implemented in the Saber circuit simulator. The major contributions made in this dissertation are

- (1) the development of the generalized methodology for physics-based modeling of IGBTs based on two-dimensional numerical device simulations, and the methodology can be useful to model other power devices;
- (2) the formulation and solution of the steady-state and transient carrier transport problems in the LIGBT, and improvement on both LDMOST/LIGBT subcell and lateral PNP BJT/LIGBT subcell models;
- (3) the development of a parameter extraction method for the LIGBT;
- (4) the implementation of the physics-based LIGBT model into the Saber circuit simulator and the model verification for various external circuit conditions.

In Chapter 2, the generalized methodology for physics-based modeling of LIGBT is presented. The generalized methodology implies not only identification of each basic semiconductor region, but also application of semiconductor equations which represent each semiconductor region. The study gives physical insight based on extensive two-dimensional numerical device simulations and describes proper physics-based modeling techniques of IGBT. The generalized methodology for physics-based modeling developed in Chapter 2 can be useful

to model the power devices with wide and lightly doped regions such as the power bipolar transistor and the PIN diode.

In Chapter 3, a physical model for the LIGBT is described. Improvements on both LDMOST/LIGBT and lateral PNP BJT/LIGBT subcell models have resulted in accurate simulation results without parameter optimization. In the LDMOST/LIGBT subcell, we have modeled the current-induced-space-charge region shown to be of great importance for improving the accuracy of the LDMOST model. In the lateral PNP BJT/LIGBT subcell, we have derived analytical geometrical factors based on the ambipolar transport equation. The lateral and quasi-vertical geometrical factors enhance the analysis of the two-dimensional anode current, and thus, result in accurate modeling for steady-state and transient conditions for the LIGBT. The implementation of the Kirk-effect into the LIGBT model also improved the model accuracy of steady-state and transient conditions. PISCES simulations and measurements of test devices support the LIGBT model, which is demonstrated in steady-state and turn-off transient Saber simulations.

In Chapter 4, the parameter extraction method for the LIGBT from terminal electrical measurements has been developed and described. Unlike microelectronic devices, the dynamic characteristic must be used to extract model parameters in the LIGBT. This is because the LIGBT has the characteristics of the internal LDMOST and lateral PNP bipolar transistor subcells. The dynamic waveforms contain many features that isolate different model parameters. The

model parameters are obtained sequentially by selecting features of the device characteristics that isolate parameters, and using the parameters obtained from previous extraction steps to calculate the model parameters from the next measured characteristics.

In Chapter 5, the methodology for implementing the physical model for the LIGBT into the Saber circuit simulator is described. The Saber is a very powerful modeling tool for power devices which have widely varying cell structures even within one category of device. The numerical machine in the Saber is transparent to the user, and one template containing the model equations is the only implementation needed. The techniques for implementing the LIGBT model equations into the Saber circuit simulator are essential for accurate simulations. Furthermore, the techniques demonstrated for modeling the LIGBT with the Saber circuit simulator are generally applicable to modeling other power semiconductor devices. The LIGBT model is verified for various external circuit conditions and for the full range of steady-state and dynamic conditions in which the LIGBT is intended to be operated. The LIGBT Saber model also is shown to be suitable for simulating the behavior of the LIGBT for general purpose external circuit conditions and designing the various power circuits.

In Chapter 6, the main accomplishments of this dissertation are summarized, and suggestions for further research are discussed.

Appendix, related to Chapter 5, lists the source code of the LIGBT Saber model template.

CHAPTER 2 GENERALIZED METHODOLOGY FOR PHYSICS-BASED MODELING OF IGBTs

2.1 Introduction

In the 1980s, the Insulated Gate Bipolar Transistor (IGBT) was introduced to overcome the high on-state power loss of the power MOSFET while maintaining the simple gate drive requirements of the device. Because of the special structure of the IGBTs, the standard models designed for microelectronic devices are not suitable to describe the dynamic behavior of the IGBTs. Unlike low voltage IC devices, IGBTs have wide and lightly doped base regions to support high-voltage operation. This lightly doped high-level injection region significantly impacts the electrical behavior of the IGBTs, and is difficult to model because its distributed phenomenon must be described by the ambipolar transport equation. This is the main reason that computer-aided-design for IGBTs is hardly used in the power electronic field today. For the design and analysis of power ASICs, we need to develop accurate models for the IGBT and implement the models into the circuit simulators like SABER and SPICE.

In this chapter, we present the generalized methodology of physics-based modeling for IGBTs. It starts from identification of the basic semiconductor regions in IGBT's mode of operation. The basic

semiconductor regions can be categorized into five parts in IGBTs: the anode-base and the base-cathode p/n junction region, the anode and cathode terminal doping region, the high-level injection base ambipolar region, the current-induced-space-charge region at the edge of the LDMOST channel, and the channel region. Each basic semiconductor region can be identified based on the two-dimensional PISCES simulation. As a matter of fact, numerous works about regional modeling of IGBTs [GOE94, ALL94, MA94] have been proposed while no works focused on the generalized methodology of modeling for IGBTs. The generalized methodology implies not only identification of each basic semiconductor region but also application of semiconductor equations in the basic semiconductor regions. We have four major semiconductor equations: the current density equations, the current continuity equations, Poisson's equation, and Boltzmann's relation. Different physical phenomena can be observed in different regions of the semiconductor. Therefore, some semiconductor equations can be neglected in the modeling of each semiconductor region. This can make the modeling of power devices simple and easy enough for circuit design. This generalized methodology can be applied to model the power devices that have wide and lightly doped base region such as the power bipolar transistor and the PIN diode.

Figure 2.1 summarizes the unified view of the generalized methodology for physics-based modeling for the IGBTs. The first step is to collect the device information, i.e., its geometry, layout, and dopant profile. The layout and processing information input into the

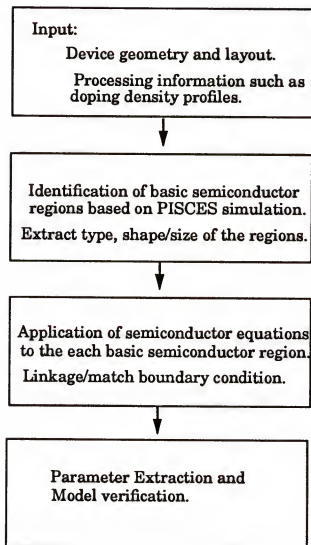


Fig.2.1 Flowchart of the generalized methodology of physics-based modeling of IGBTs.

PISCES simulator to provide insight into the device operation. Based on extensive PISCES simulations, we can identify the basic semiconductor regions in the desired modes of operation. Besides identifying the important semiconductor regions in the device, the two-dimensional PISCES simulation can be useful in determining the sizes and the shapes of each semiconductor region. Then, we can apply the semiconductor equations which represent each semiconductor region. In some regions, only one main semiconductor equation can be applied because one main physical phenomenon is observed. By solving the equations in each semiconductor region and linking them with the boundary conditions, we can make a regional model for the IGBT.

2.2 Identification of Basic Semiconductor Regions

The identification of the basic semiconductor regions is an important part of the physics-based modeling of power devices. As stated in Section 2.1, we have five different basic semiconductor regions in IGBTs. In this section, the identification of the basic semiconductor regions during the IGBT's operation is described. This study is based on extensive simulations of the IGBT using the two-dimensional device simulator PISCES, which utilizes finite-element approximations and a Gummel-Newton numerical iteration method to solve the discretized semiconductor equations defining the carrier transport [COU88]. Doping density profiles and structural dimensions for our test devices were obtained from AT&T Bell laboratories. The

simulations reveal the internal mechanisms of IGBT such as different collecting regions of the holes and the electrons, and they show that the current-induced-space-charge region forms at the edge of the LDMOST channel of the LIGBT because of its similar structure to the RESURF LDMOST. Only a depletion region is observed in the base area of the VIGBT [HEF88]. Hence, the current transport for the LIGBT and the VIGBT are very different from each other.

The PISCES input file for the LIGBT is listed in the Appendix A. The geometrical dimensions and the doping density profiles for each region are specified in the input file. This device was fabricated with a dielectric isolation (DI) Bipolar-CMOS-DMOS technology developed at AT&T Bell Laboratories [KOS93]. Figure 2.2 shows the cross-section of the LIGBT generated by the PISCES input file. From the doping profiles, the LIGBT has five different doped regions: anode, deep p^+ , p^- body, source, and the n^- uniform doped base, and it has three electrodes: anode, cathode, and gate.

As shown in Fig.2.2, the LIGBT has two subcells indicating an LDMOS gate structure with the bipolar current conduction of the lateral PNP BJT. In this structure, the current is negligible when a negative voltage is applied to the anode with respect to the cathode because the anode-base junction will become reverse-biased. This provides the LIGBT with its reverse blocking capability. When a positive voltage is applied to the anode terminal with the gate short-circuited to the cathode terminal, the base-cathode junction becomes reverse-biased, and the device operates in its forward blocking mode. At posi-

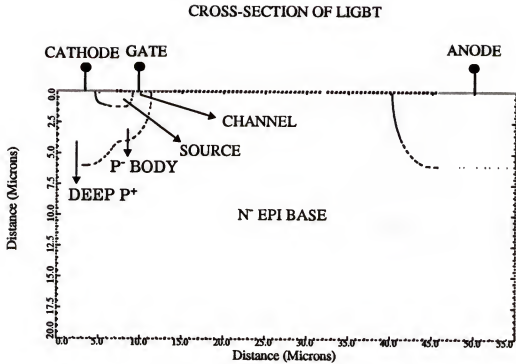
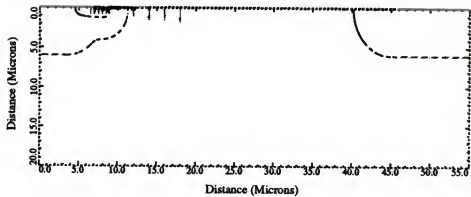


Fig.2.2 Cross-section of the AT&T LIGHT (not scaled). Five different doped regions and three electrodes: anode, cathode, and gate.

tive anode terminal voltages, if a positive gate bias is applied of sufficient magnitude to invert the surface of the p^- body region under the gate (channel), the LIGBT operates in its forward conducting state because electrons can now flow from the n^+ source region to the n^- epi base region. This is well shown in Fig. 2.3(a). The electron current vector appears only in the source and the channel regions. In this forward conducting state, the anode-base junction is forward-biased and the anode p^+ region injects holes into the n^- epi base region shown in Fig. 2.3(b). The hole current vector predominates throughout the entire LIGBT structure (except in the source region) in the forward conducting mode.

The LIGBT has a LDMOST/LIGBT subcell and a lateral PNP BJT/LIGBT subcell. When the forward bias (anode voltage) is increased, the injected hole concentration increases until it exceeds the background doping level of the n^- base region. When the excess carrier density exceeds the background doping level by several orders of magnitude, as is the case in the injected part of the base region, the high-injection condition is satisfied. The transport of electrons and the transport of holes are coupled by the electric field in the drift region, and they cannot be treated separately. Consequently, the basic equation governing the carrier transport is the ambipolar transport equation. Figure 2.4(b) shows the one-dimensional carrier concentration at $y=2\mu\text{m}$, and it clearly shows that the doping concentration of the injected carriers are much higher than the base background doping concentration ($=2 \times 10^{14} \text{cm}^{-3}$) and the electron and hole concentrations

(a)

LIGBT Electron Current Vector at $V_a=10$ V $V_g=10$ V

(b)

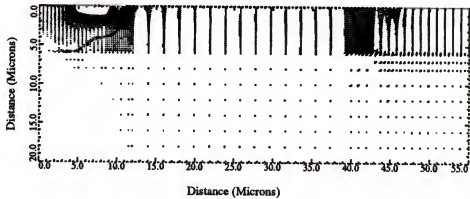
LIGBT Hole Current Vector at $V_a=10$ V $V_g=10$ V

Fig.2.3 (a) Electron and (b) Hole current vectors of LIGBT at $V_a=10$ V and $V_g=10$ V.

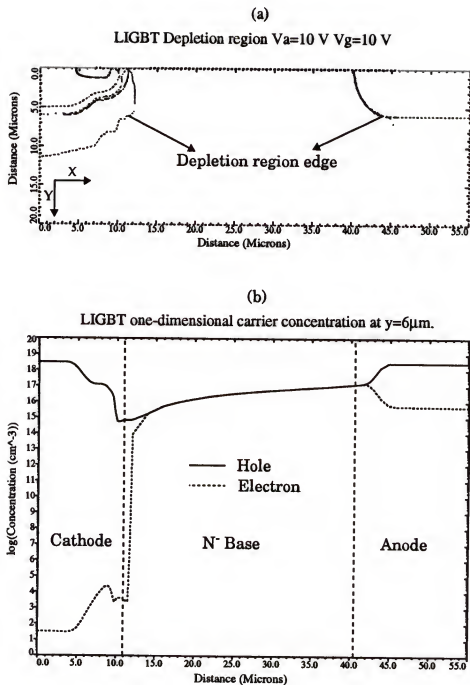


Fig.2.4 (a) Depletion region in the LIGBT at $V_a=10\text{ V}$ and $V_g=10\text{ V}$.
 (b) One dimensional carrier distribution in the base at $y=6\mu\text{m}$.

in the base region is equal ($n = p$). From this, we know the base region is a high-level-injected ambipolar transport region.

The LDMOST/LIGBT subcell has the structure of the RESURF LDMOST. The RESURF LDMOST has the n -epi layer which functions as a drift region. The thickness and doping density of the n -epi layer are controlled to produce a low surface electric field to provide high breakdown voltage [APP79]. At high currents, excess electrons accumulate in the drift region due to drift-velocity saturation. The excess electrons produce space-charge-limited current flow and compensate for the widening of the depletion region of base-collector junction in the lateral PNP BJT/LIGBT subcell. Figure 2.5(a) shows the equipotential distribution of two-dimensional PISCES simulation of the LIGBT mode of operation at $V_{GS} = 10$ V and $V_{AK} = 10$ V. The dotted line indicates the edge of the base-collector depletion region. As shown by PISCES simulation, the equipotential contour not only is distributed along the base-collector depletion region but also is distributed along the quasi-neutral base region near the channel. The equipotential contour along the quasi-neutral base has a cone-shaped, cross-sectional area. The cone-shaped area is a current-induced-space-charge (CISC) region which is incorporated into the base current. Figure 2.5(b) shows the PISCES simulation of a one-dimensional electric field distribution at the surface ($y=0.001\mu\text{m}$). In Fig.2.5(a), the physical location of the drain of LDMOST/LIGBT subcell is shown.

The simulated PISCES current flowlines within the LIGBT device is shown in Fig.2.6. This simulated current flowlines reflect

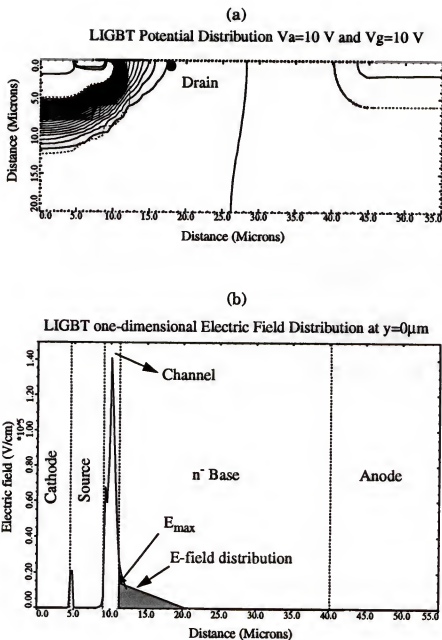


Fig.2.5 (a) Equipotential contour from two-dimensional PISCES simulation of the LIGBT mode of operation.
 (b) One-dimensional electric field distribution at the surface.

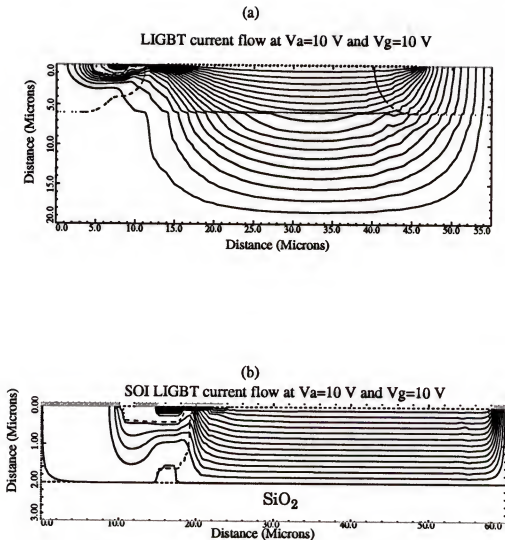


Fig.2.6 Anode current flowlines from two-dimensional PISCES simulation at $V_{GS}=10$ V and $V_{AK}=10$ V. The flowlines are plotted such that the equal currents ($\Delta I = 0.2 \times 10^{-3}$ A/ μm) flow between any two adjacent flowlines (a) LIGBT (scaled) (b) SOI LIGBT (not scaled).

operation below the onset of latchup. Electrons injected (predominantly laterally) from the source drive the lateral PNP BJT/LIGBT base harder, producing additional forward bias on the anode junction and additional hole injection into the n^- base. This enhances the conductivity modulation of the lateral PNP BJT/LIGBT base region, including the adjacent depletion region at the lateral PNP BJT/LIGBT base-collector junction. Normal LIGBT operation is evident; the LDMOST current through the channel supplies electrons to the base of the lateral PNP/BJT subcell which then supports, through recombination, the hole transport current. The LIGBT anode current is the sum of the LDMOST channel current and the collector current of the lateral PNP BJT/LIGBT subcell. Note that all the current flowlines are not lateral, but some have quasi-vertical flowlines. The SOI LIGBT [DIS93], in Fig. 2.6(b), has predominantly lateral flowlines. This is mainly because the SOI LIGBT has only $2\mu\text{m}$ of n^- epi base region. The geometrical parameters such as base width, junction depth, and base length influence the current flowlines. Hence, we need to model two-dimensional analysis of current flowlines which are dependent on the geometrical input parameters.

Now that the identification of the basic semiconductor regions in the LIGBT is completed, and the circuit models for each building blocks can be defined. Figure 2.7 shows the block diagram of the basic semiconductor regions in the LIGBT mode of operation. It consists of seven blocks, and each block represents a basic semiconductor region. Among the basic semiconductor regions, the high-level injection ambi-

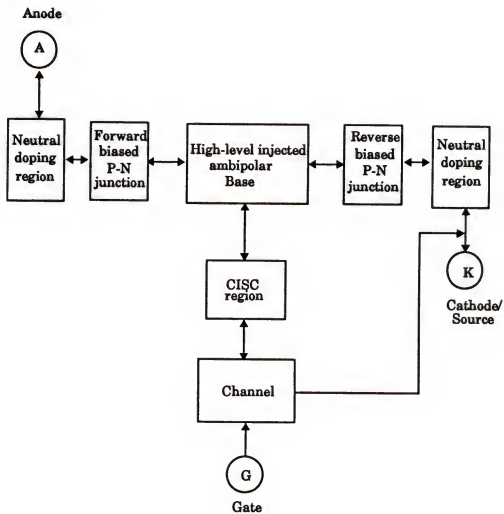


Fig.2.7 Block diagram of basic semiconductor regions for the LIGBT mode of operation

polar base region is the most important region because it determines the electrical behavior of the LIGBT. In next section, the application of semiconductor equations to the basic semiconductor region is illustrated.

The PISCES input file, which details the geometry input and processing information, is listed in Appendix A.

2.3 Application of Semiconductor Equations

As stated in Section 2.2, the LIGBT exhibits distinctive basic semiconductor regions, such as the lightly doped ($\sim 10^{14} \text{ cm}^{-3}$) base region for blocking high voltages where most of the excess carriers are stored, the more heavily doped ($\sim 10^{20} \text{ cm}^{-3}$) end regions for the terminals, and the lightly doped ($\sim 10^{16} \text{ cm}^{-3}$) semiconductor surface in an MOS structure. The semiconductor equations - the current density equations, the current continuity equations, Poisson's equation, and Boltzmann's equation- do not have the same importance in the different regions of the IGBTs. Some of the semiconductor equations can be neglected in modeling each semiconductor region. In Table 2.1, the basic semiconductor regions with identification criteria and the semiconductor equations which fit each region are listed. Except the channel region, all other regions are related to the behavior of the bipolar power semiconductor devices.

TABLE 2.1 THE MAIN TYPES OF BEHAVIOR INSIDE IGBT'S

Region	Identification criteria	Equations
P/N junction region	Seperation of quasi-fermi level in P/N junction under forward and reverse biased condition. Carrier depletion in reverse biased condition.	Boltzmann and Poisson equations
Neutral ohmic region for terminal	Quasi-neutral doped region for terminal with forward -biased P/N junction	Diffusion current equation
High-level-injected ambipolar region	Equal majority and minority carriers which are higher than the background doping	Ambipolar transport equation
Current-induced-space-charge region	majority carrier accumulation in the region	Poisson Equation
Channel region	MOSFET structure	MOS current equation

2.3.1 P/N Junction Region

Using Boltzmann statistics, the carrier concentration p_0 at the forward biased p⁺/n junction is simply given as

$$p_0 = \left(\frac{n_i^2}{N_{epi}} \right) e^{\left(\frac{qV_{EB}}{KT} \right)} \quad (2.1)$$

where n_i is the silicon intrinsic carrier concentration, N_{epi} is the doping concentration in the base, and V_{EB} is the voltage drop across the forward junction.

Poisson's equation is used to describe the variation of junction depletion width with voltage

$$\frac{d^2 V}{dx^2} = \frac{\rho(x)}{\epsilon_{si}} \quad (2.2)$$

where ϵ_{si} is the silicon dielectric constant and $\rho(x)$ is the charge density. The depleted region of the p⁺n junction extends significantly into the n⁻ region during reverse bias, squeezing the holes and electron charges into a smaller volume. The positive charges in the depleted region consist of immobile ionized donors and injected holes from the base. Integrating (2.2) twice over the depleted region and rearranging the equation gives

$$x_d = \left(\frac{2\epsilon_{si}(V_{BC} + \phi_{bi})}{qN_{epi}} \right)^{\frac{1}{2}} \quad (2.3)$$

where x_d is the depletion region width, ϕ_{bi} is the built-in potential of base-collector junction, and V_{BC} is the voltage drop across the reverse-biased junction.

2.3.2 Doping Region for Terminals

When the doping concentration of neutral ohmic region for terminal is high ($\sim 10^{20} \text{ cm}^{-3}$) as in anode for the IGBT, the minority carrier (electron) from the epi-base is back-injected to the anode. This back-injection current is part of anode current and it can not be neglected. The transport is mainly by minority carrier diffusion current. The electron current with negligible recombination in the p^+/n forward-biased space-charge region is defined by the recombination in the quasi-neutral p^+ anode region as [ERN91]

$$J_{BA} = j_{N0} p_0^2 \quad (2.4)$$

where $j_{N0} = \frac{qD_n}{L_n N_A}$, D_n is the electron diffusion coefficient, L_n is the electron diffusion length, and N_A is the acceptor doping concentration in the anode.

2.3.3 High-Level-Injected Ambipolar Region

In general, the electron and hole currents are given by

$$I_n = nq\mu_n AE + qAD_n \frac{\partial n}{\partial x} \quad (2.5)$$

$$I_p = pq\mu_p AE - qAD_p \frac{\partial p}{\partial x} \quad (2.6)$$

where I_n, I_p are the electron and hole currents, respectively, n, p are the electron and hole carrier concentrations, respectively, A is the device active area, q is the electronic charge, E is the electric field, D_n, D_p are the electron and hole diffusivities, respectively. The first terms in Eqns (2.1) and (2.2) are due to drift and the second terms are due to diffusion. Under the high-gain conditions of the conventional bipolar transistor analysis, these equations can be decoupled and the transport of minority carriers in the base for both high- and low-level injection conditions.

However, for the low-gain, high-level injection conditions of the lateral PNP BJT/LIGBT subcell, the difference between the electron drift and diffusion currents is significant and the net electron current can not be set to zero in approximating the electric field. Since the net electron current has a significant effect on the hole drift current, the electron and hole transport equations can not be decoupled. Assuming quasi-neutrality and high-level injection in the base region, the current equations can be written for the ambipolar transport case by eliminating the electric field between Eqns (2.5) and (2.6) [HEF86]:

$$I_n = \frac{b}{1+b} I_T + qD_A A \frac{\partial n}{\partial x} \quad (2.7)$$

$$I_p = \frac{1}{1+b} I_T - qD_A A \frac{\partial p}{\partial x} \quad (2.8)$$

where $b = \frac{\mu_n}{\mu_p}$ is the electron/hole mobility ratio, I_T is the total current, D_A is the ambipolar diffusivity, which is $2D_nD_p/(D_n+D_p)$. Equations (2.3) and (2.4) depend on the total current so that the transport of electrons and holes are coupled.

The continuity equation for hole is

$$\frac{\partial p}{\partial t} = -\frac{p}{\tau_H} - \frac{1}{q} \frac{\partial J_p}{\partial x} \quad (2.9)$$

where τ_H is the high-level injection lifetime. From Eqn (2.8) and (2.9) the time-dependent ambipolar diffusion equation is obtained

$$\frac{\partial^2 p}{\partial x^2} = \frac{p}{L_A^2} + \frac{1}{D_A} \frac{\partial p}{\partial t} \quad (2.10)$$

where L_A is the ambipolar diffusion length and is $\sqrt{D_A \tau_H}$, τ_H is the high-level injection lifetime.

A requirement in deriving this expression is that the total current I_T (anode current) is independent of position in the base. This is satisfied in the LIGBT because the base current (electrons) flows from the collector through the base in the same direction as the injected hole current from the emitter. Equation (2.10) does not hold for the traditional bipolar transistor for which the base current enters from the side in the middle of the base.

2.3.4 Current-Induced-Space-Charge Region

Poisson's equation can be used to describe the voltage across the CISC region

$$\frac{\partial^2 V}{\partial x^2} = -\frac{\rho(x)}{\epsilon_s} \equiv \frac{q}{\epsilon_s} (N_{EPI} - n) \quad (2.11)$$

where N_{EPI} is the uniform doping concentration of the epi-base and the n is the electron density in the CISC region.

Because of relatively low doping in epi-base region, when MOS current becomes higher than the critical current, $J_C = qN_{EPI}V_{sat}$, the CISC region forms. Because the MOS current is velocity saturated in the CISC region, the MOS current is constant with x through the entire CISC region. This is based on the one-dimensional high current Kirk effect [KIR62] in the bipolar transistor. Because the Kirk effect of bipolar transistor is based on the one-dimensional analysis, there can be a error to apply the one-dimensional model into the two-dimensional CISC region in the LIGBT. However, PISCES simulation results (Fig. 2.5) show the our one-dimensional CISC region model can be applied in the LIGBT CISC region.

When MOS current is less than J_C , there is no CISC region near channel. Therefore, V_{CH} equals to V_{DS} .

From Fig. 2.5(b), the voltage drop across the CISC region as a function of x can be assumed to be the grey triangular area based on the PISCES simulation, and its expression is

$$V_{CISC} = \frac{1}{2} E_{max} W_{cisc} \quad (2.12)$$

where W_{cisc} is the width of the CISC region and is assumed to be equal to the base-collector depletion region width based on the PISCES sim-

ulation, E_{max} is the peak electric field in CISC region occurring at the edge of the LDMOST channel.

Poisson's equation for the CISC region, defining $x=0$ as the channel edge of the base and $x=W_{cisc}$ as the CISC region edge of the quasi-neutral base, takes the form

$$\frac{dE}{dx} = \frac{q}{\epsilon_{si}} \left(N_{epi} - \frac{J_{MOS}}{qv_{sat}} \right) \quad (2.13)$$

and

$$\int_{E_{max}}^0 dE = \frac{q}{\epsilon_{si}} \left(N_{epi} - \frac{J_{MOS}}{qv_{sat}} \right) \int_0^{W_{cisc}} dx \quad (2.14)$$

From Eqn.(2.14), E_{max} is

$$E_{max} = -\frac{q}{\epsilon_{si}} \left(N_{epi} - \frac{J_{MOS}}{qv_{sat}} \right) W_{cisc} \quad (2.15)$$

Using Eqns (2.12) and (2.14), the final voltage drop across the CISC region can be expressed as

$$V_{CISC} = \frac{q}{2\epsilon_{si}} \left(\frac{J_{MOS}}{qv_{sat}} - N_{epi} \right) W_{cisc}^2 \quad (2.16)$$

where J_{MOS} is the current density of the MOS current and v_{sat} is the electron saturation velocity.

2.3.5 Channel Region

The MOS current of the n-channel LDMOST is a nonlinear function of V_{CH} and the gate to source voltage drop V_{GS} . As the chan-

nels of LIGBT structures may be short ($<2\mu\text{m}$), current saturation due to velocity saturation rather than vanishing inversion charge and channel length reduction during current saturation have to be taken into account. Because the transconductance parameter is different for the linear and the saturation region, both linear and saturation transconductance parameters must be extracted.

The gate-to-source capacitance C_{gs} , and the gate-to-drain capacitance C_{gd} are both nonlinear functions of V_{GD} and V_{GS} . On the one hand, they show the typical MOS capacitance behavior; on the other hand, charge sharing and charge shielding phenomena giving rise to a further degree of complexity and modeling effort have to be taken into account in order to describe these capacitances accurately. The general expressions which describe the current and the charges are presented in Section 3.2.2 and Section 3.3.1.

2.4 Summary

The physical insights with regard to two structures of IGBTs (LIGBT and SOI LIGBT) have been obtained by using the two-dimensional device simulator PISCES. These insights were used to identify five different basic semiconductor regions. The application of correct semiconductor equations to each basic semiconductor region is critical in modeling the power devices.

From the PISCES simulations, we found that IGBTs have five different basic semiconductor regions: the p/n junction region, the end-

doping region for the terminals, the high-level-injected ambipolar region, the current-induced-space-charge region, and the channel region. The high-level injection region inside the IGBT reduces the on-state device voltage drop but also increases the turn-off transient. The high-level injection region is the area of carrier storage that is a distributed phenomenon, described by the ambipolar transport equation. Therefore, the high-level-injected ambipolar region is one of the most important regions in modeling the IGBT. The modeling of power devices with a high-level-injection ambipolar region can be greatly simplified using our modeling technique. The generalized methodology for the physics-based modeling of IGBT developed herein can be useful in modeling power devices with wide and lightly doped regions such as the power bipolar transistor and the PIN diode.

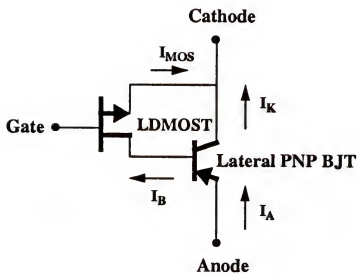
CHAPTER 3 CELL MODEL DEVELOPMENT FOR LIGBT

3.1 Introduction

This chapter presents a physical model for the Lateral Insulated Gate Bipolar Transistor (LIGBT). The LIGBT is a high-voltage device that can be integrated through both junction and dielectric isolation techniques. The LIGBT is a MOS-controlled power switching device for power integrated circuit applications. The schematic structure of the device is shown in Fig.1.1. The $n^{\text{-epi}}$ is lightly doped base region to support high voltages. The p^+ ($\sim 10^{18} \text{ cm}^{-3}$) cathode region has a higher doping density than the p -body ($\sim 10^{16} \text{ cm}^{-3}$) region and is included mainly to improve static latch-up. Fig.3.1 also shows the LIGBT is dielectrically isolated in a silicon tub from the substrate. Thus we have not modeled the parasitic PNP transistor formed between the anode and the p -type substrate in the conventional device.

The operation of the LIGBT simply can be treated as a partitioning of an n -channel LDMOST (LDMOST/LIGBT subcell) and a lateral PNP BJT (lateral PNP BJT/LIGBT subcell). The basic symbolic circuit of the LIGBT is shown in Fig.3.1(a). The LIGBT functions as a bipolar transistor that is supplied base current by a MOSFET [Yil84].

(a)



(b)

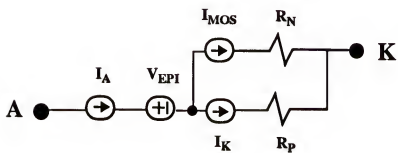


Fig.3.1 (a) The symbolic circuit of the LIGBT.

(b) The basic equivalent circuit of the LIGBT.

Contrasted to VLSI bipolar transistors, the lateral PNP BJT/LIGBT subcell has a low base lifetime, a low current gain and is operated mainly in high-level injection in the base region. Since high-level injection prevails ambipolar theory must be used to describe the transport of electrons and holes in the n^- epi base region. The modeling of LDMOST/LIGBT subcell is also important because it provides the input base current to the lateral PNP BJT/LIGBT subcell.

Although previous works have modeled the operation of Lights [FOS88, PAT86, ERN91], none has specifically modeled the current-induced-space-charge (CISS) region in the n^- epi base region near the channel. The CISC region has a cone-shaped, cross-sectional area which is incorporated into the base current. The partitioning of the lateral PNP BJT/LIGBT subcell anode current into a lateral and a quasi-vertical part is also important and has not been modeled before.

We have improved both subcell models by considering the above two criteria. Modeling the CISC region in the relatively low-doped base region results in a more accurate LDMOST model. Analytical geometrical factors [CHO92] based on the ambipolar transport equation for the lateral and the quasi-vertical anode currents for the lateral PNP BJT/LIGBT subcell have been derived. The total anode current is computed from two one-dimensional currents. For the quasi-vertical anode component, a geometrical factor corrects for its two dimensionality.

The model is implemented in the Saber circuit simulator [SAB92]. Saber is a very powerful modeling tool for power devices

which have widely varying cell structures even within one category of device (for example, LDMOST). The numerical "machine" in Saber is transparent to the user. One template containing the model equations is all the software that needs implementation. This is not true of SPICE typically where, aside from the model subroutine, 17 other sub-routines have to be modified.

3.2 LIGBT Steady-State Model

In the on-state under steady-state conduction, the LIGBT cathode electrode is grounded and the anode electrode is connected through a load to a positive bias supply. When the gate voltage is higher than the threshold voltage, the LDMOST channel is inverted. The LDMOST current constitutes the base current of the lateral PNP BJT/LIGBT subcell. As the emitter-base junction turns on, it injects holes into the epi base region and thus modulates its conductivity. The injected hole current can be divided into two components. The first component is a recombination current with LDMOST current injected in the epi base and anode regions. The second component is collected by the p^+ cathode that forms the collector of the lateral PNP BJT/LIGBT subcell. The steady-state operation of the Lights can be generally described with a simple equivalent circuit, which is shown in Fig.3.1(b) [SIM85, PAT86, KIM88]. The two-dimensional model of anode current for the lateral PNP BJT/LIGBT subcell and the model of the CISC region for the LDMOST/LIGBT subcell are added to the sim-

ple model to improve the accuracy of the circuit simulator for the LIGBT devices.

A system of parametric equations is derived for the steady-state electron- and hole-current densities and the excess carrier concentration and the emitter-base voltage in next subsection. These equations are obtained by solving the ambipolar transport equations for the boundary conditions of the lateral PNP BJT/LIGBT subcell. The characteristics of the lateral PNP BJT/LIGBT subcell are then combined with the LDMOST/LIGBT subcell model to completely describe the steady-state current-voltage characteristics of the LIGBT.

3.2.1 Lateral PNP BJT/LIGBT Subcell Model

The analysis is performed using the coordinate system defined in Fig.3.2. Defining $x = 0$ as the emitter edge of the base and $x = W$ as the collector edge of the quasi-neutral base, the steady-state boundary conditions for the hole distribution are

$$p(W) = 0, p(0) = p_0$$

Solving the steady-state ambipolar diffusion (Eqn(2.10) with $\frac{\partial p}{\partial t} = 0$) in the base region with the above boundary conditions yields one-dimensional hole distribution [FOS86]

$$p(x) = p_0 \frac{\sinh\left(\frac{W-x}{L_A}\right)}{\sinh\left(\frac{W}{L_A}\right)} \quad (3.1)$$

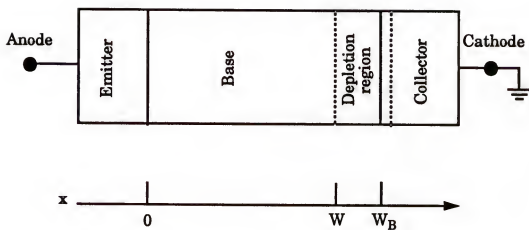


Fig.3.2 Coordinate system used in developing the LIGBT model. The junction space-charge region is indicated by the dashed lines.

where p_0 is the hole density at the edge of the emitter-base junction space-charge region.

From the lateral PNP BJT/LIGBT subcell, the anode-cathode voltage (V_{AK}) can be divided into three components: (1) a p^+/n^- forward-biased emitter-base junction voltage (V_{EB}), (2) a n^-/p reverse-biased base-collector junction voltage (V_{BC}), and (3) V_{EPI} the voltage drop across the epi base region. Thus,

$$V_{AK} = V_{EB} + V_{BC} + V_{EPI} \quad (3.2)$$

Using quasi-equilibrium assumption [MAC87], p_0 can be expressed in terms of V_{EB}

$$p_0 = \frac{n_i^2}{N_{epi}} e^{\left(\frac{qV_{EB}}{KT}\right)} \quad (3.3)$$

where n_i is the intrinsic carrier concentration, K is the Boltzmann constant, and T is the temperature. W can be calculated through the depletion approximation at the reverse-biased collector-base junction:

$$W = W_B - \left(\frac{2\epsilon_{si}(V_{BC} + \phi_{bi})}{qN_{epi}} \right)^{\frac{1}{2}} \quad (3.4)$$

where W_B is the metallurgical base width, ϕ_{bi} is the built-in potential of base-collector junction, N_{epi} is the uniform doping density in the base region, and ϵ_{si} is the dielectric constant of silicon. The base-collector reverse-biased voltage, V_{BC} will shrink the width W of the quasi-

neutral base. In addition, however, W is affected by the current density, which flows through the depletion layer between base and collector. This phenomenon is called Kirk effect [KIR62] in the theory of the bipolar transistor. In the LIGBT structure, the base-collector depletion layer spreads predominantly into the lightly doped base, and the high collector current tends to increase the width of the quasi-neutral base region. Consequently, the width of quasi-neutral base becomes

$$W = W_B - \left[\frac{2\epsilon_{si}(\phi_{bi} + V_{BC})}{qN_{epi}} \right]^{\frac{1}{2}} \left(1 + \frac{J_K}{J_C} \right)^{-\frac{1}{2}} \quad (3.5)$$

where J_K is the collector current density of the lateral PNP BJT/LIGBT subcell, and $J_C = qN_{epi}v_s$, critical current, where v_s denotes the hole saturation velocity. As clear from Eqn.(3.5), a high collector current will increase the width of the quasi-neutral base, and thus the Kirk effect opposes the formation of the depletion region occurring because of the reverse bias across the base-collector junction.

Now we can determine V_{EPI} with $p \approx n$:

$$J_p = q\mu_p \left(p(x)E - \frac{KTdp(x)}{qdx} \right) \quad (3.6)$$

$$J_n = q\mu_n \left(p(x)E + \frac{KTdp(x)}{qdx} \right) \quad (3.7)$$

and total current density, J_A , is

$$J_A = J_n + J_p \quad (3.8)$$

Combining Eqn.(3.6) and Eqn.(3.7) gives

$$E = \frac{J_A}{q(\mu_n + \mu_p)} \frac{1}{p(x)} + \frac{KT(b-1)}{q(b+1)} \frac{dp(x)}{dx} \quad (3.9)$$

$$V_{EPI} = \int_0^{W_M} E dx \quad (3.10)$$

where W_M is defined at the point where $p(x)$ equals N_{epi} .

Finally V_{EPI} is

$$V_{EPI} = \frac{J_A L_A \sinh\left(\frac{W}{L_A}\right)}{q p_0 (\mu_n + \mu_p)} \left[\ln\left(\tanh\left(\frac{W}{2L_A}\right)\right) - \ln\left(\tanh\left(\frac{W - W_M}{2L_A}\right)\right) \right] \\ + \frac{KT(b-1)}{q(b+1)} \ln\left(\frac{p_0}{N_{epi}}\right) \quad (3.11)$$

The first of these terms is due to an ohmic drop in the epi base region, whereas the second is caused by the asymmetric concentration gradient produced by the unequal electron and hole mobilities [GHA77].

The anode current is a sum of the hole current, which is injected from emitter (anode) into the epi base, and the electron current, which is back injected into the emitter from the epi base. For negligible recombination in the emitter-base junction space-charge region, J_A can be related to p_0 following conventional p-i-n diode theory [GHA77]

$$J_A = j_{N_0} p_0^2 - q D_p \left. \frac{dp(x)}{dx} \right|_{x=0}$$

$$= j_{N0} p_0^2 + \frac{q D_n p_0}{L_A} \coth\left(\frac{W}{L_A}\right) \quad (3.12)$$

where $j_{N0} = \frac{q D_n}{L_n N_A}$, D_n is the electron diffusion coefficient, L_n is the electron diffusion length, and N_A is the acceptor doping concentration in the anode.

Under steady-state on condition, LDMOST current flows into the epi base region through the LDMOST channel. This LDMOST current is equal to the base current of the lateral PNP BJT/LIGBT subcell. The base current is the sum of the recombination current in the epi-base region and the electron current, which is back injected into the emitter from the epi base. The latter contribution is the same as the first term in Eqn.(3.12)

$$\begin{aligned} J_B &= j_{N0} p_0^2 + q \int_0^W \frac{p(x)}{\tau_H} dx \\ &= \frac{q p_0 L_A}{\tau_H} \left[\coth\left(\frac{W}{L_A}\right) - \operatorname{csch}\left(\frac{W}{L_A}\right) \right] \end{aligned} \quad (3.13)$$

The collector current density J_K is expressed as $J_A - J_B$. Previously, the cross sectional area A taken from layout has been used to relate the terminal currents I_A and I_B (and I_K) to the respective current densities. This is true for one-dimensional current analysis and the one-dimensional current analysis has been used to model the LIGBT devices. However, it has been reported [LIN67] that lateral devices with wide base such as lateral PNP transistor do not have one-

dimensional current flow but rather have a two-dimensional current flow. To gain some insights about the current flowline for the LIGBT, a 2-D numerical device simulator, PISCES has been used (see Fig.2.6).

The PISCES simulation implies the correctness in modeling the anode current flow as two components, a purely lateral current component I_{AL} in the anode side-wall region and a quasi-vertical component along the curved trajectory I_{AV} .

The one-dimensional anode current based on the ambipolar transport equation can be written as

$$I_{A0} = \left(\frac{qD_p p_0}{L_A} \coth\left(\frac{W}{L_A}\right) + j_{n0} p_0^2 \right) p_a x_j \quad (3.14)$$

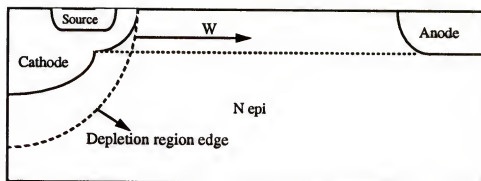
where, and $p_a x_j$ is the area of the anode junction sidewall, these geometrical parameters are graphically shown in Fig. 4.1.

The purely lateral current which flows parallel to the surface can be modeled as a function of the varying base width due to lateral diffusion in the anode. The geometry of the anode sidewall is shown in Fig.3.3. The curved junctions on the anode side considered to be cylindrical with radius of curvature x_j , the junction depth. The base width W_b is a function of y

$$W_b(y) = W + x_j(1 - \cos\theta) \quad (3.15)$$

$$\text{where } \theta = \sin^{-1}\left(\frac{y}{x_j}\right)$$

(a)



(b)

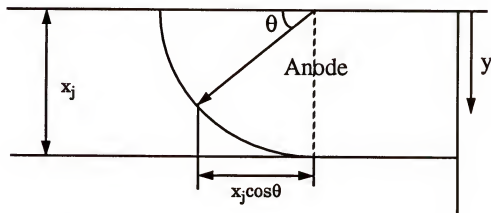


Fig.3.3 (a) The geometry of anode and cathode sidewall of the LIGBT. The base width W_b is a function of x , W is a base width, and x_j is the junction depth. (b) close-up of geometry of anode sidewall.

Since the lateral current component is assumed purely horizontal to the surface

$$\begin{aligned}
 I_{AL} &= p_a \int_0^{x_j} |J_A| dx \\
 &= \frac{qD_p p_0 p_a}{L_A} \int_0^{x_j} \coth \left(\frac{W_b(y)}{L_A} \right) dx + j_{N0} p_0^2 p_a x_j
 \end{aligned} \tag{3.16}$$

where

$$\int_0^{x_j} \coth \left(\frac{W_b(y)}{L_A} \right) dx = \int_0^{\frac{\pi}{2}} \cos \theta \coth \left(\frac{W + x_j (1 - \cos \theta)}{L_A} \right) d\theta \tag{3.17}$$

The lateral geometrical factor F_{GAL} is

$$F_{GAL} = \frac{I_{AL}}{I_{A0}} \tag{3.18}$$

and it only depends on the base width W , the junction depth x_j , and the ambipolar diffusion length L_A .

The vertical anode current path can be modeled by concentric curves with radius r as shown in Fig.3.5. This model is valid if the maximum base width W_{bm} of the device, shown in Fig. 4.1, is at least twice the length of the device thickness D_{VT} as in our case. However, if D_{VT} is approximately equal to W_{bm} , there is at least a 30% error in the vertical geometrical factor. Most of conventional LIGBT power device

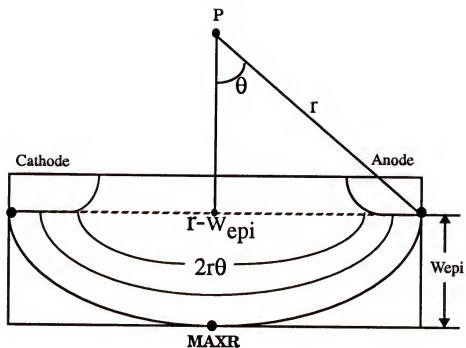


Fig.3.4 The geometry of the concentric-curve model. The half-circle lines indicate the quasi-vertical component of anode current.

have $W_{bm} \gg D_{VT}$, thereby validating this concentric-curve model for the LIGBT.

The concentric path of quasi-vertical anode current can be seen and verified from two-dimensional PISCES simulation as shown in Fig. 2.6. To model the two-dimensional anode current effects in the lateral PNP BJT/LIGBT subcell, the derivation of the quasi-vertical geometrical factor is needed. Without this geometrical factor, the anode current should be assumed to flow laterally through the entire base region in the LIGBT. This assumption may result the overestimation of the anode current because more current density can be found in the lateral current flow region (because of short distance between the anode and the cathode). Figure 3.13 shows the comparison of transient simulation results between our model and the model without the two-dimensional current effect.

From Fig.3.4, the length of the curvature is $2r\theta$. Then the quasi-vertical geometrical factor F_{GAV} is

$$F_{GAV} = \frac{W}{x_j} \int_{MINR}^{MAXR} \frac{1}{2r\theta} dr \quad (3.19)$$

and

$$\cos \theta = \frac{r - W_{epi}}{r} \quad (3.20)$$

where MAXR is the maximum radius and MINR is the minimum radius of the curvature. Major parameters, taken from the layout and cross-sectional geometry, are the W_{bm} , the epitaxial thickness W_{epi} ,

and the junction depth x_j . If we assume the circle of the MAXR is crossing three points as shown in Fig.3.4, then we can find the center point of the circle. Finally, MAXR is

$$MAXR = \sqrt{\left(\frac{\left(\frac{W_{bm}}{2}\right)^2 - W_{epi}^2}{2W_{epi}}\right)^2 + \left(\frac{W_{bm}}{2}\right)^2} \quad (3.21)$$

and MINR is

$$MINR = MAXR - W_{epi} \quad (3.22)$$

Then, Eqn.(3.19) becomes

$$\begin{aligned} F_{GAV} &= \frac{W}{2x_j} \int_a^b \frac{\sin \theta}{\theta (1 - \cos \theta)} d\theta \\ &= \frac{W}{2x_j} \left[2\theta^{-1} + \frac{1}{6}\theta + \frac{1}{1080}\theta^3 + \frac{1}{7600}\theta^5 + \dots \right]_b^a \end{aligned} \quad (3.23)$$

where $a = \text{acos}\left(1 - \frac{W_{epi}}{MAXR - W_{epi}}\right)$ and $b = \text{acos}\left(1 - \frac{MAXR}{W_{epi}}\right)$.

The total geometrical factor F_{GA} is the sum of F_{GAL} and F_{GAV} . Assuming no variation in V_{EB} along the emitter-base junction, the complete analytical expression for the anode current is

$$I_A = I_{A0} F_{GA} \quad (3.24)$$

where $F_{GA} = F_{GAL} + F_{GAV}$

The ratio of F_{GAL} and F_{GAV} will be used later to calculate the high-level-injection quasi-vertical base region charges for transient analysis.

3.2.2 LDMOST/LIGBT Subcell Model

When the gate voltage is raised above the threshold voltage of the LDMOST gate, the LDMOST channel region is inverted. As the anode voltage increases, the n-channel LDMOST is turned-on and electron current starts to flow into the base region through LDMOST channel. The LDMOST current is the input base current of the lateral PNP BJT/LIGBT subcell.

In the LIGBT normal mode of operation, when the LDMOST current is higher than the critical current, $I_c = qAN_{epi}v_{sat}$ (a typical LIGBT operating condition because of the relatively low doping in n⁻ epi base region), the current-induced-space charge (CISC) region forms in the epi-base region near the channel. The voltage across the CISC region is

$$V_{CISC} = \frac{1}{2}E_{max}W_{cisc} \quad (3.25)$$

where W_{cisc} is the width of the CISC region and is assumed to be equal the base-collector depletion region width, E_{max} is the peak electric field in CISC region occurring at the edge of the LDMOST channel. Then E_{max} becomes

$$E_{max} = \frac{q}{\epsilon_{si}} \left(\frac{J_{MOS}}{qV_{sat}} - N_{epi} \right) W_{cisc} \quad (3.26)$$

The final form of V_{CISC} becomes

$$V_{CISC} = \frac{q}{2\epsilon_{si}} \left(\frac{J_{MOS}}{qV_{sat}} - N_{epi} \right) W_{cisc}^2 \quad (3.27)$$

As shown in Fig.2.5(a), the drain is located in the base region near the channel and drain-source voltage V_{DS} in LDMOST/LIGBT subcell is approximately V_{BC} in the lateral PNP BJT/LIGBT subcell, and the channel voltage V_{CH} is different from V_{DS} . The drain-source voltage is a sum of the channel voltage and the voltage drop across the CISC region.

$$V_{CH} = V_{DS} - V_{CISC} \quad (3.28)$$

Equation (3.26) explains why V_{CH} , not V_{DS} which is typically used, must be used to calculate an accurate value of LDMOST current.

We used the first order MOSFET current equation in our LDMOST/LIGBT subcell model for its simplicity. It is suggested that more refined LDMOST current equations may be needed in the LDMOST/LIGBT subcell model. When the LIGBT is in the on-state, the LDMOST is in its linear region and the MOSFET first order linear region equation is given by

$$I_{MOS}^{lin} = K_{Plin} (V_{GS} - V_{TH}) V_{CH} - \frac{K_{Plin}}{2} V_{CH}^2 \quad (3.29)$$

where $K_{P_{lin}}$, transconductance parameter for linear region, is equal to the product of the oxide capacitance, the surface mobility, and the effective width-to-length ratio of the LDMOST/LIGBT subcell. One of the unique properties of the IGBT is that the transconductance parameter is different for the linear and the saturation region.

The LIGBT is in its current saturation region when the LDMOST is in its saturation region. The first order saturation current equation of the MOSFET is given by

$$I_{MOS}^{sat} = \frac{K_{P_{sat}}}{2} (V_{GS} - V_{TH})^2 \quad (3.30)$$

where $K_{P_{sat}}$ is the transconductance parameter for saturation region, and its extraction method will be illustrated in Chapter 4. The saturation is assumed to be due to velocity saturation at the end of the channel for simplicity.

Figure 3.5 shows the measured and simulated current-voltage characteristics with a gate voltage of 8 V. In the figure, the vertical axis shows the anode current and the horizontal axis shows the anode-cathode voltage. Also, the I-V characteristics for simulations of the new model using V_{CH} and the old model using V_{DS} are compared. Clearly, the new model provides a better fit to the measurement than the old model. The new model shows a saturation in V_{CH} which is supported by PISCES simulations. Above $V_{AK} = 1.0$ V, the old model severely overestimates the anode current because V_{DS} continues to increase monotonically as anode-cathode voltage increases.

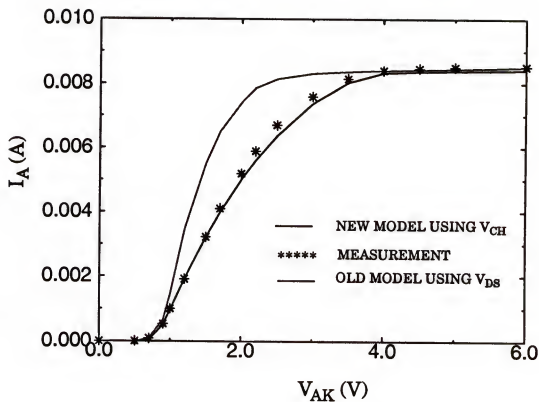


Fig.3.5 I-V characteristics comparison between new model using V_{CH} and old model [FOS88] using V_{DS} , contrasted with measurement at $V_{GS} = 8.0$.

3.2.3 LIGBT Steady-State Model-Routine

The flowchart for semi-numerical LIGBT steady-state model-routine is given in Fig. 3.6. The model has three internal nodes: V_{EB} , V_{BC} , and V_{CH} . To solve for the internal node voltages, three implicit bi-secant semi-numerical loops have been implemented in FORTRAN code.

The model routine, given V_{AK} , V_{GS} , first determines the node voltages of the lateral PNP BJT/LIGBT subcell (V_{EB} , V_{BC}) using the quasi-static equations developed in Section 3.2.1. V_{BC} is assumed to be V_{DS} in the LDMOST/LIGBT subcell; V_{CH} can be determined by iteration until the LDMOST current equals the base current of the lateral PNP BJT/LIGBT subcell. Finally, we can determine V_{EB} using the bi-secant method.

3.3 LIGBT Transient Model

In this section, the turn-off process in the LIGBT with a resistive load is discussed in reference to the switching mode of operation. After removal of the gate voltage, the LDMOST current falls rapidly, removing the base current to the lateral PNP BJT/LIGBT subcell. However, the collector current of the lateral PNP BJT/LIGBT subcell falls more slowly because of the predominance of the excess carrier recombination in epi base region. Figure 3.7 shows a circuit configuration used to measure the dynamic characteristics. The measurement of anode current turn-off waveform of the LIGBT is shown in Fig. 3.8.

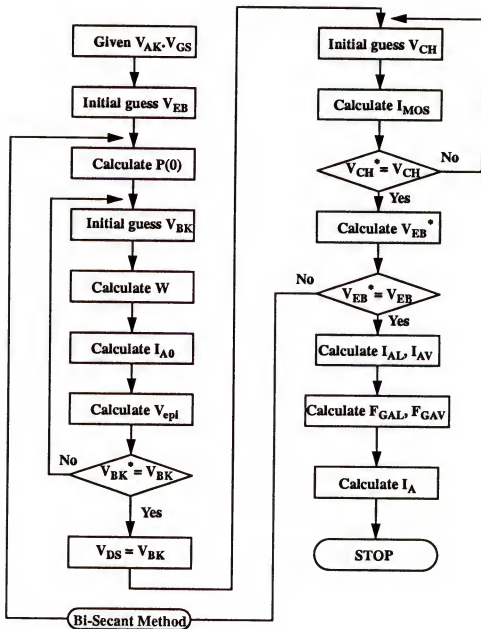


Fig.3.6 The LIGBT steady-state model flowchart.

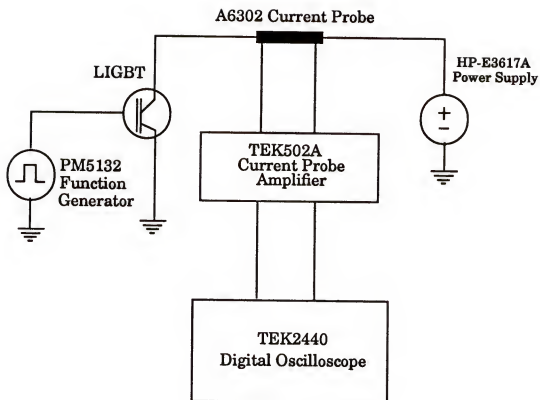


Fig.3.7 Circuit configuration used to measure the dynamic characteristics.

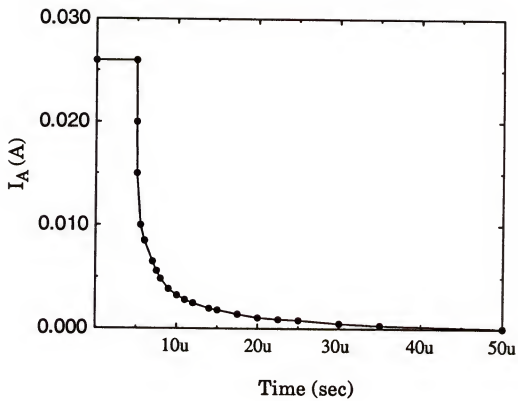


Fig.3.8 Measured LIGBT anode current versus time for turn-off transient.

The current decay waveform has three phases. The sudden drop that results from cessation of the LDMOST electron current is the first phase. Following that, the current decay is governed by carrier recombination in the conductivity-modulated lightly-doped epi-base region. The decay is slow because the ambipolar high-level-injection lifetime is long. As the level of injected carriers decrease below the doping level, the current decay then is dictated by the low-level-injection lifetime which is shorter than that of the high-injection case.

3.3.1 Lateral PNP BJT/LIGBT Subcell Model

Once the LDMOST current is removed, the stored excess majority carriers (electron) in the base region decay by recombination and by injection into the anode. The excess of minority carriers (hole) in the base continues to be supplied by injection from anode and is depleted by collection at the base-collector junction and recombination in the base in such a way that quasi-neutrality is maintained throughout the base. The collected hole current is equal to the lateral PNP BJT/LIGBT subcell current during the slowly decaying current phase at a constant voltage shown in Fig. 3.8.

The predominant charges stored in the lateral PNP BJT/LIGBT subcell are Q_{BE} , Q_{BC} , and Q_{JC} . Q_{BE} and Q_{BC} are the hole charges stored in the quasi-neutral base of the lateral PNP BJT/LIGBT subcell associated with the anode and cathode, respectively. In general, Q_{BE} and Q_{BC} depend on both the V_{EB} and V_{BC} . This means that the charging current $\frac{dQ_{BC}}{dt}$ has two terms

$$\frac{dQ_{BC}}{dt} = \left(\frac{\partial Q_{BC}}{\partial V_{BC}} \right) \frac{dV_{BC}}{dt} + \left(\frac{\partial Q_{BC}}{\partial V_{BE}} \right) \frac{dV_{BE}}{dt} \quad (3.31)$$

The first term can be represented as a capacitor, and the second term as a transcapacitance. Without the $\frac{dQ_{BC}}{dt}$ term in the LIGBT transient model, the simulation of turn-off transient will be in significant error to the results because $\frac{dQ_{BC}}{dt}$ reflects the finite carrier transit time in the quasi-neutral base.

Since the injected base region charge is a function of anode current, each Q_{BE} and Q_{BC} should be partitioned into two components, lateral, and quasi-vertical. The partitioning is defined by integrating the carrier continuity equations in the base. This is physically representative of the bipolar transistor [FOS86]

$$Q_{BEL} = qD_{vw}x_j \int_0^W \left(1 - \frac{x}{W} \right) p(x, t) dx \quad (3.32)$$

$$Q_{BCL} = qD_{vw}x_j \int_0^W \left(\frac{x}{W} \right) p(x, t) dx \quad (3.33)$$

where Q_{BEL} and Q_{BCL} are the lateral components of the base charge associated with anode and cathode, respectively, x_j is the junction depth, and D_{vw} is the device width, shown in Fig. 4.1.

From Eqn. (3.1)

$$p(x, t) = p(0, t) \left[\frac{\sinh \left[\frac{(W-x)}{L_A} \right]}{\sinh \left(\frac{W}{L_A} \right)} \right] \quad (3.34)$$

Inserting Eqn. (3.32) into Eqn. (3.30) and Eqn. (3.31) yields

$$Q_{BEL} = qD_{vW}x_jL_Ap(0,t) \left[\coth\left(\frac{W}{L_A}\right) - \frac{L_A}{W} \right] \quad (3.35)$$

and

$$Q_{BCL} = qD_{vW}x_jL_Ap(0,t) \left[\frac{L_A}{W} - \operatorname{csch}\left(\frac{W}{L_A}\right) \right] \quad (3.36)$$

The time dependence of Q_{BEL} and Q_{BCL} are defined by the time dependences of the emitter-base and base-collector voltages which control $p(0,t)$ and W .

The geometrical factors can be used to calculate Q_{BEV} and Q_{BCV} , the quasi-vertical components of the base charge associated with anode and cathode, respectively

$$Q_{BEV} = \frac{F_{GAV}}{F_{GAL}} Q_{BEL} \quad (3.37)$$

$$Q_{BCV} = \frac{F_{GAV}}{F_{GAL}} Q_{BCL} \quad (3.38)$$

Since Q_{BE} and Q_{BC} are the sum of its lateral and quasi-vertical components

$$Q_{BE} = Q_{BEL} + Q_{BEV} \quad (3.39)$$

$$Q_{BC} = Q_{BCL} + Q_{BCV} \quad (3.40)$$

Q_{JC} is the reverse-biased base-collector depletion region charge

$$Q_{JC} = qAN_{epi}x_d \quad (3.41)$$

where A is the cross-sectional area, x_d is the base-collector depletion region width.

3.3.2 LDMOST/LIGBT Subcell Model

The charges in the LDMOST/LIGBT subcell are Q_{GS} , Q_{GD} , and Q_{CISC} . In our LDMOST charge model we did not include the channel charge in Q_{GS} and Q_{GD} for simplicity. It is suggested that our LDMOST charge model should be refined to improve the accuracy of the transient analysis.

Q_{GS} is the gate oxide capacitance of the source overlap charge

$$Q_{GS} = L_{sol}D_{VW}C_{ox}V_{GS} \quad (3.42)$$

where L_{sol} is the length of the gate oxide overlapping in source.

Q_{GD} is gate-drain capacitance times gate-drain voltage

$$Q_{GD} = C_{GD}V_{GD} \quad (3.43)$$

C_{GD} is equal to the gate oxide capacitance of the gate-drain overlap for $V_{DS} \leq V_{GS}$, but for $V_{DS} > V_{GS}$ the silicon beneath the gate-drain overlap becomes depleted and the gate-drain capacitance consists of the series combination of the gate-drain overlap oxide capacitance and the gate-drain overlap depletion capacitance.

$$C_{GD} = \begin{cases} C_{oxd} & \text{for } (V_{DS} \leq V_{GS}) \\ \frac{C_{oxd}C_{gdj}}{(C_{oxd} + C_{gdj})} & \text{for } (V_{DS} > V_{GS}) \end{cases} \quad (3.44)$$

where $C_{oxd} = L_{dol}D_{vW}C_{ox}$ is gate oxide capacitance of the drain overlap charge and $C_{gdj} = L_{dol}D_{vW}\epsilon_{si}/W_{gdj}$ is the gate-drain depletion capacitance, L_{dol} is the length of the gate oxide overlapping in drain, W_{gdj} , the gate-drain depletion width, is proportional to the square-root of the gate-drain voltage.

Unlike the VIGBT [HEF90], the silicon beneath the gate-drain overlap not only becomes depleted but also becomes a excess electron accumulation region in the LIGBT. The accumulation region is called the current-induced-space-charge (CISC) region. After the gate is turned-off, the charges in CISC region dissipate by recombination in the base region. The charge in the CISC region is

$$Q_{CISC} = qA_{cisc}D_{vW}\left(\frac{J_{MOS}}{qV_{sat}} - N_{epi}\right) \quad (3.45)$$

where $A_{cisc} = \frac{1}{4}\pi W_{cisc}^2$ is the cross-sectional area of CISC region.

3.3.3 LIGBT Network Representation

The complete network representation of the new LIGBT model is shown in Fig. 3.9. The model is not a simple equivalent circuit. The elements shown represent physical components of current and voltage in the LDMOST, described implicitly by the analysis, which requires a numerical solution.

The transient behavior of the LIGBT is simulated using both the quasi-static currents/voltages and charging currents $\frac{dQ}{dt}$. The

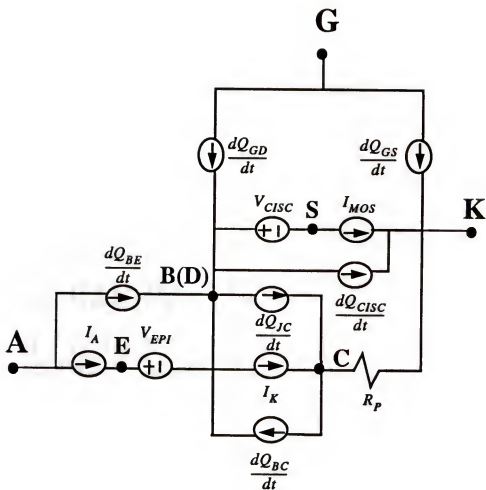


Fig.3.9 Complete network representation of the charge-based LIGBT model.

charges within the device are expressed in terms of node voltages, and the transient charging currents are represented by time derivatives of the charges.

3.4 LIGBT Model Verification

Measured and simulated current-voltage characteristics for three different values of gate voltages are shown in Fig. 3.10. The measured data in Fig.3.10 were taken with an HP-4145 parameter analyzer. The following features of the static characteristics are indicated in Fig. 3.10: 1) the diode voltage offset due to the anode-epitaxial layer of p/n junction at low V_{AK} (less than $\sim 0.7V$), 2) the on-state region which has low-resistance due to conductivity modulation, 3) the current saturation region due to saturation of the LDMOST channel current, and 4) the triode-like region which is referred to as the linear region for LDMOST. The agreement between measurement and simulation for 1) the diode voltage offset validates the implementation of V_{EB} , 2) the on-state characteristics validates the implementation of V_{EPI} , 3) the saturation characteristics validates the implementation of I_{MOS} , I_K , and I_A , and 4) the triode-like region validates the implementation of V_{CH} . The agreement is good even without the optimization of the parameters.

Figure 3.11 compares the channel voltages V_{CH} from PISCES simulation and our model as a function of V_{AK} , contrasted with the characteristic for V_{DS} . Because the value of V_{DS} is assumed to be

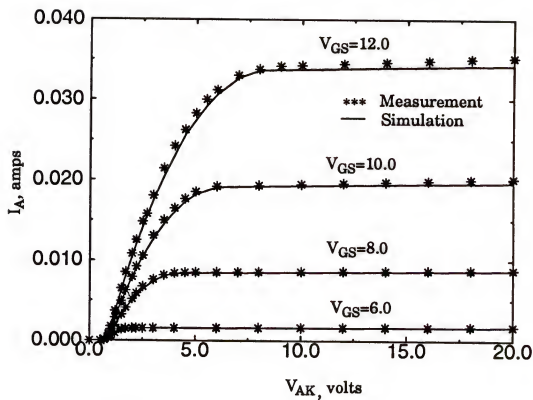


Fig.3.10 Measured and simulated I-V characteristics for four different values of gate-source voltages ($V_{GS}=6.0, 8.0, 10.0$, and 12.0 volts)

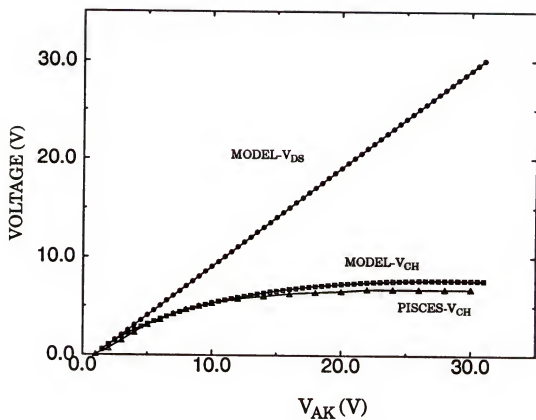


Fig.3.11 PISCES and model simulated V_{CH} versus V_{AK} , contrasted with V_{DS} at $V_{GS} = 10.0$ volts.

equal to the value of V_{BK} , V_{DS} increases monotonically as V_{AK} increases. We used V_{DS} in our model and obtained poor agreement with the measurement data in Fig. 3.9, which leads us to believe the model optimization is suspected in earlier models [FOS88, PAT86, ERN91]. We emphasize we did not perform any optimization of the model parameters for the simulation in Fig. 3.9. Both PISCES and our model simulations show the saturation of V_{CH} as V_{AK} increases. It is obvious that a more accurate value of the LDMOST current can be obtained by using V_{CH} instead of V_{DS} . Since the LDMOST current is the input base current of the lateral PNP BJT/LIGBT subcell and of comparable magnitude to anode current, an accurate value for the LDMOST current is essential in predicting an accurate value for anode current.

The two-dimensional model of anode current also improves the accuracy of our LIGBT model. The lateral and quasi-vertical geometrical factors vary dependent on bias conditions. At $V_{AK} = 3$ V and $V_{GS} = 10$ V, F_{GAL} is 1.42 and F_{GAV} is 0.84. In the conventional one-dimensional model, the anode current was calculated simply as the anode current density times area. Thus, it would require optimization in the parameters for the conventional one-dimensional model to fit the simulation result to the measurement.

In Fig. 3.12, a constant-anode-supply-voltage test circuit is used to verify the transient model. Measured and simulated transient turn-off characteristics of the LIGBT are compared in Fig. 3.13. The turn-off transient simulation shows good agreement with measure-

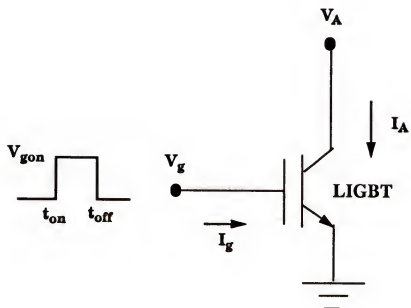


Fig.3.12 Circuit configuration of an LIGBT with a constant anode voltage.

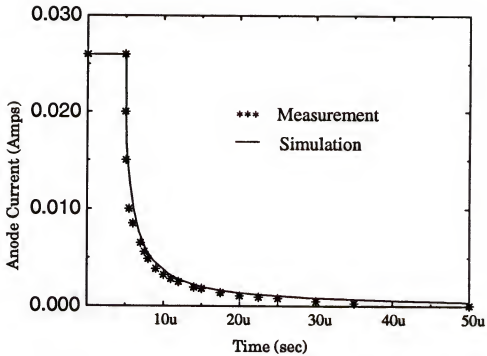


Fig.3.13 Measured and simulated LIGBT anode current versus time for turn-off transient at $V_{GS}=10$ V and $V_{AK}= 2.0$ V.

ment. The high-level injection lifetime of $5.6\mu\text{sec}$, the threshold voltage of 3.4volts , and device width of $120\mu\text{m}$ were used to simulate the transient analysis. The high-level injection lifetime was calculated from the transient measurement (Fig. 3.8) and the parameter extracting method is well explained in Chapter 4. The good fit between the simulation and the measurement on 1) initial rapid drop validates the implementation of LDMOST current I_{MOS} , and 2) the slow decay of the turn-off current waveform validates the model for Q_{bt} and Q_{bv} , the lateral and the quasi-vertical two-dimensional base region charges, and the partitioned charges, Q_{BE} and Q_{BC} .

In Fig. 3.13 we have the simulated result (solid curve) which have been obtained using our model. So here the CISC region model and the two-dimensional current model have been included in the calculations. The dashed curve represent the simulated result, which is obtained without the CISC region model and the two-dimensional current model, but the other details of the theory have been unchanged.

There are two distinctive differences between two curves. First is the total current magnitude difference. As stated in Section 3.2.2, the anode current can be overestimated without using the CISC region model. Without the two-dimensional current model the anode current also can be overestimated because the anode current is assumed to flow linearly in whole area of the device. Second is the difference of turn-off decay time. The turn-off decay time of the solid curve ($\sim 7\mu\text{s}$) is much faster than the turn-off decay time of the dashed curve ($\sim 12\mu\text{s}$). The turn-off time of $4\mu\text{s}$ is more reasonable value than

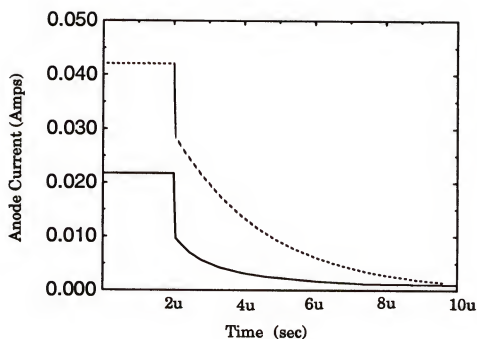


Fig.3.14 Turn-off transient anode current simulation comparison between our LIGBT model (solid curve) and the LIGBT model without the CISC region model and the two-dimensional current model (dashed curve).

the turn-off time of $7\mu\text{s}$ for a high-level injection lifetime of $5.6\mu\text{s}$ because of the significant carrier removal via recombination in the p+ anode. The dashed curve has longer turn-off decay time due to its one-dimensional current model. Without two-dimensional current model, the base charge is calculated from the integration of hole charge from 0 to w (base width) times device area. This can lead the overestimation of the base charge.

3.5 Summary

We have developed an accurate steady-state and transient model for LIGBT. Improvements on both subcell models of the LIGBT have resulted in accurate simulation results without parameter optimization. In the LDMOST/LIGBT subcell, we have modeled the CISC region based on PISCES simulation. The model of CISC region has been shown to be of great importance for improving the accuracy of the LDMOST model. In the lateral PNP BJT/LIGBT subcell, we have derived analytical geometrical factors based on the ambipolar transport equation. The lateral and quasi-vertical geometrical factors enhance the analysis of the two-dimensional anode current, and, thus, result in the accurate modeling for steady-state and transient conditions for the LIGBT. The results of the simulation show good fit with both steady-state and transient measurements without the optimization of the model parameters. The LIGBT model has been used to

simulate various external circuits and has been proven valid for simulating various power circuits. This will be shown in Chapter 5.

CHAPTER 4 PARAMETER EXTRACTION FOR IGBT

4.1 Introduction

Contrasted to microelectronic devices where the steady-state current-voltage characteristics in conjunction with interelectrode capacitance-voltage characteristics are sufficient to extract most of the device model parameters, the dynamic characteristics must in general be examined to characterize the devices and to extract model parameters for merged power devices such as the IGBT. This is because the characteristics of the internal MOSFETs and bipolar transistors are convoluted in the steady-state characteristics. The dynamic measurements require high precision and the ability to make calculations on waveforms obtained from measurements. In addition, calculations are required to separate the measured characteristics into the characteristics of the internal MOSFETs and bipolar transistor.

In our IGBT model, parameters can be divided into two major categories. The first category is device parameters. The device parameters are layout dependent and thus can differ from device to device in the same IC. The IGBT model has nine device parameters defined in Table 4.1 and Fig. 4.1. The second category is model parame

Table 4.1 LIGBT DEVICE PARAMETERS

Name	Description	Units	Default
D_{VW}	Device Width	m	240e-6
D_{VT}	Device Thickness	m	18.0e-6
X_J	Junction Depth	m	6.5e-6
P_A	Anode Perimeter Length	m	18.0e-6
L_{AJ}	Anode Implantation Window Length	m	6.0e-6
W_{BM}	Maximum Base Width	m	65.0e-6
W_{B0}	Minimum Base Width	m	50.0e-6
L_{SOL}	Gate Oxide Source Overlapping	m	1.0e-6
L_{DOL}	Gate Oxide Drain Overlapping	m	4.0e-6

Table 4.2 LIGBT MODEL PARAMETERS

Name	Description	Units	Default
τ_H	High-Injection Lifetime	sec	2.4e-6
V_{TH}	Threshold Voltage	V	4.8
N_{EPI}	Uniform Base Doping	m^{-3}	1.0e21
A_{DP}	Peak Anode Doping	m^{-3}	1.0e25
P_{BD}	P-Body Doping	m^{-3}	5.0e22
$K_{P_{sat}}$	Saturation Transconductance Parameter	A/V^2	4.1e-4
$K_{P_{lin}}$	Linear Transconductance Parameter	A/V^2	9.6e-4

ters. The model parameters are process-dependent and generally common to all the LIGBT devices on the same chip. The complete set of seven parameters is shown in Table 4.2. The key model parameters are τ_H , V_{TH} , $K_{P_{sat}}$, and $K_{P_{lin}}$, and can be obtained by DC and turn-off transient measurement. The others, N_{EPI} , A_{DP} and P_{BD} , are technological, and can be obtained from processing information.

The PISCES simulation results are used to extract the parameters because we could not get the good measurement data of transient analysis. Three basic types of simulations are used to extract the key model parameters: 1) The turn-off current tail decay rate versus anode current characteristic is used to extract the high-injection lifetime τ_H . 2) The saturation current versus gate voltage characteristic is used to extract the MOSFET-channel saturation-region transconductance parameter $K_{P_{sat}}$ and MOSFET-channel threshold-voltage V_{TH} . 3) The on-state voltage versus gate voltage characteristic at a constant anode current is used to extract the MOSFET-channel linear-region transconductance parameter $K_{P_{lin}}$. The three simulations are performed in the indicated sequence because the extracted parameters taken from simulated data require the values of the previously extracted parameters.

4.2 High-Level-Injection Lifetime

The simulated turn-off transient anode current of the LIGBT with the constant anode supply voltage test circuit is shown in Fig.

4.2. Two different phases of the turn-off current are shown. An initial fast drop that results from the termination of LDMOST channel current followed by slow decay due to the recombination of excess holes injected by the anode into the base region. The slow decay also has two phases, high-level-injection lifetime decay and low-level-injection lifetime decay. During the first phases of slow decay, the current decay is dictated by the high-level injection lifetime. As the level of injected carriers decreases below the base doping-level, the lifetime decreases because the base enters the low-level injection condition. Therefore, during the second phase of the slow decay, the current decay is dictated by the low-injection lifetime which is shorter than that for the high-injection case [DAR87].

The first phase of the slow decay has been matched to an exponential waveform with 99.998% of accuracy. If we assume the first phase of the slow decay has an exponential form, then the anode current is

$$I(t) = I_C \exp\left(-\frac{t}{\tau_H}\right) \quad (4.1)$$

where I_C is the initial value of anode current of the first phase of the slow decay.

$$\ln\left(\frac{I(t)}{I_C}\right) = -\frac{t}{\tau_H} \quad (4.2)$$

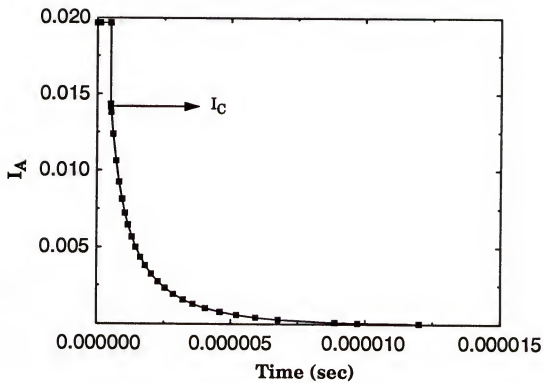


Fig.4.2 Simulated turn-off transient anode current versus time with the constant anode supply voltage test circuit.

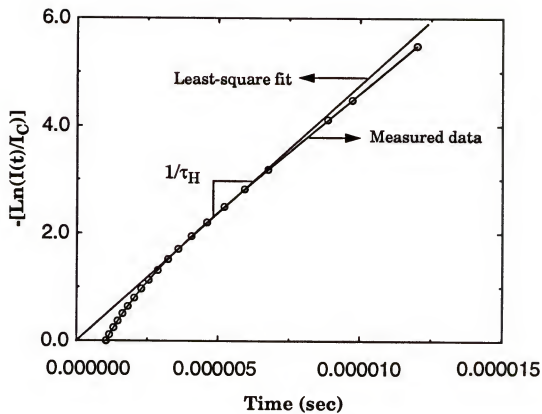


Fig.4.3 The anode current tail decay rate indicating that the slope of the least-square fit is equal to the $1/\tau_H$.

Figure 4.2 shows the simulated values of the turn-off transient of the anode current versus time, and a least-squares fit to the anode current decay. The value of high-level injection lifetime, τ_H , can be extracted from the slope of the least-square fit of Eqn.(4.2). The graphic output of this extraction routine is given in Fig.4.3 and the slope is equal to the $1/\tau_H$.

4.3 Threshold Voltage and Saturation Transconductance

Figure 4.4 shows a simulated constant-voltage turn-off transient current waveform with four different gate-source voltages at a constant anode voltage. The characteristic consists of an initial rapid fall followed by a slowly decaying current tail. Because the source of the majority carriers (the LDMOST current) is absent during the slowly decaying phase of the current waveform, the total stored excess majority carriers in the base decays by recombination and by injection into the anode. The excess majority carriers in the base continue to be supplied by injection from the anode and are depleted by collection at the base-collector junction and recombination in the base in such a way that quasi-neutrality is maintained throughout the base. The collected minority carrier (hole) current is equal to the total LIGBT current during the slowly decaying current phase at a constant voltage. It is shown that the initial rapid fall in current is not equal to the base current (LDMOST current) for two reasons: 1) the anode voltage and the quasi-neutral base width are in general different during the

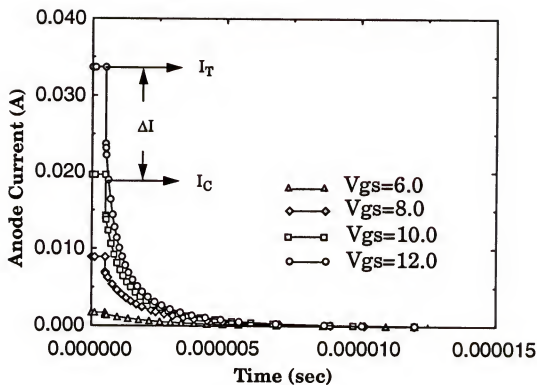


Fig.4.4 Simulated constant voltage turn-off current waveform with four different gate-source voltages at a constant anode voltage $V_{AK}=2.0$.

collector current decay than during steady-state [HEF88], and 2) removal of the electron current also reduces the hole current. This indicates that the quasi-static approximation is not valid for the devices governed by ambipolar transport equation. Hence, the initial drop of the forward current ΔI is usually less than the LDMOST channel current due to the compensation made by the base-collector junction displacement current. However, the effect of those compensation currents can be minimized provided that we limit V_{CE} to a relatively small value by selecting a small resistor in our measurement [SHE93]. Therefore, ΔI can be approximately considered as I_{MOS} . The common-collector current gain β is given by

$$\beta = \left. \frac{I_C}{I_T - I_C} \right|_{V_A = \text{constant}} \quad (4.3)$$

When the LDMOST current I_{MOS} enters the current saturation regime, the base current of the lateral PNP BJT/LIGBT subcell becomes constant. Therefore, the LIGBT anode current saturates when the internal LDMOST current saturates where the LDMOST saturation current is magnified $(1+\beta)$ by the steady-state common collector current gain of the lateral PNP BJT/LIGBT subcell. Fig.4.5 shows the lateral PNP BJT/LIGBT subcell common-collector current gain β versus anode current, which was extracted from the dynamic characteristics of the LIGBT. At high currents, the value of β decreases due to the increased rate of electron-current injection into the anode. Once β is known, the saturation characteristics of the inter-

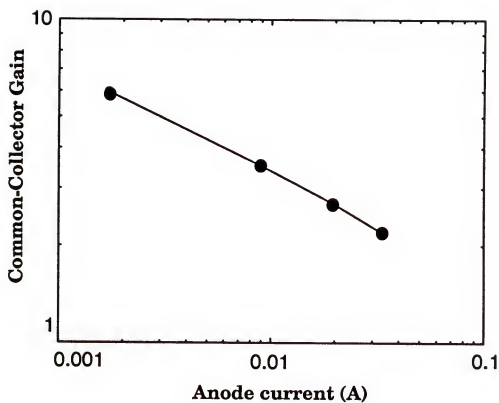


Fig.4.5 The common collector current gain versus anode current for the lateral PNP BJT/LIGBT subcell of LIGBT.

nal LDMOST are obtained from LIGBT saturation characteristics by dividing out the lateral PNP BJT/LIGBT current-gain

$$I_{MOS}^{sat} = \frac{I_T^{sat}}{(1 + \beta)} \quad (4.4)$$

The square root of the saturation current is linearly related to the gate-source voltage with a zero current intercept of V_{TH} and with a slope of $\sqrt{K_{Psat}/2}$ such that

$$\sqrt{I_{MOS}^{sat}} = \sqrt{\frac{K_{Psat}}{2}} (V_{GS} - V_{TH}) \quad (4.5)$$

Figure 4.6 shows the square root of the LDMOST saturation current versus gate-source voltage obtained by dividing the simulated LIGBT saturation current by the common-collector current gain. The values of K_{Psat} and V_{TH} are extracted from the slope and the zero-current intercept of this curve.

4.4 Linear Transconductance

Figure 4.7 shows the steady-state linear characteristics of the LIGBT, indicating the on-state voltage for different gate voltages at a constant anode current. From the lateral PNP BJT/LIGBT subcell, the anode-cathode voltage (V_{AK}) can be divided into three components: 1) a p^+/n^- forward-biased emitter-base junction voltage (V_{EB}), 2) a n^-/p reverse-biased base-collector junction voltage (V_{BC}), and 3) V_{EPI} the voltage drop across the epi-base region

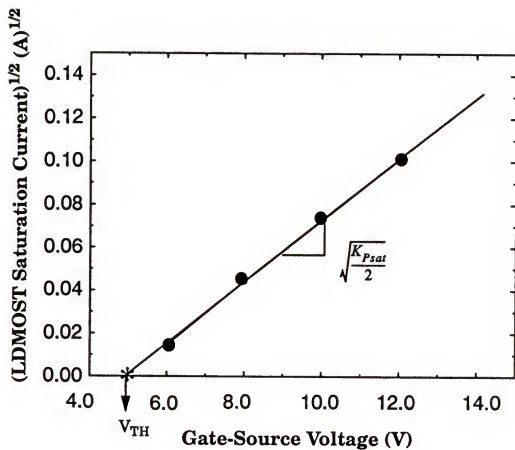


Fig.4.6 The square root of the LDMOST saturation current indicating that the slope is used to extract $K_{P_{sat}}$ and the zero current intercept is used to extract V_{TH} .

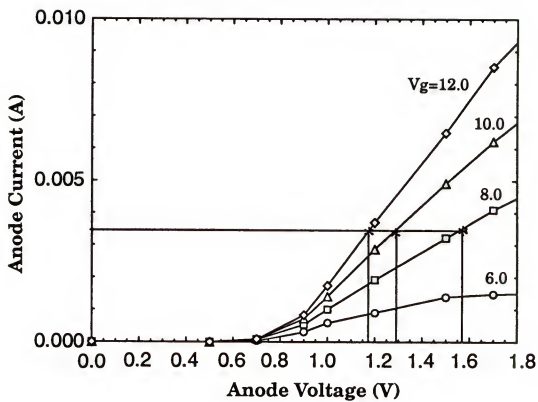


Fig.4.7 The steady-state linear characteristics of the LIGBT, indicating that the on-state voltage for different gate voltages at a constant anode current.

$$V_{AK} = V_{EB} + V_{BC} + V_{EPI} \quad (4.6)$$

For the LIGBT on-state, the LDMOST channel current is in the linear region, and this LDMOST channel current is magnified by the common collector-current gain of the lateral PNP BJT/LIGBT subcell shown in Fig. 4.5

$$I_{MOS}^{lin} = \frac{I_T^{lin}}{(1 + \beta)} \quad (4.7)$$

The drain-source voltage V_{DS} in LDMOST/LIGBT subcell is approximately V_{BC} in lateral PNP BJT/LIGBT subcell, and if it is assumed that the LDMOST channel current is given by the first term in the linear region expression of Eqn.(3.27)

$$V_{on} = V_{EB} + V_{EPI} + \frac{I_T}{(1 + \beta) K_{Plin}} \left(\frac{1}{V_{GS} - V_{TH}} \right) \quad (4.8)$$

From Eqn.(4.8), it is expected that the on-state voltage versus $1/(V_{GS} - V_{TH})$ at a constant anode current is linear with a slope of $I_T / K_{Plin}(1 + \beta)$ and with an infinite gate voltage intercept of $V_{EB} + V_{EPI}$. Fig. 4.8 shows the linear graph of anode voltage versus $1/(V_{GS} - V_{TH})$. As indicated in Fig.4.8, the value of K_{Plin} is obtained from the slope of the least-squares fit to the measured on-state voltage versus $1/(V_{GS} - V_{TH})$ at a constant anode current over the range of gate voltages where the LDMOST channel current is given by the first term in the linear region expression of Eqn.(3.27). The value of K_{Plin} is thus obtained

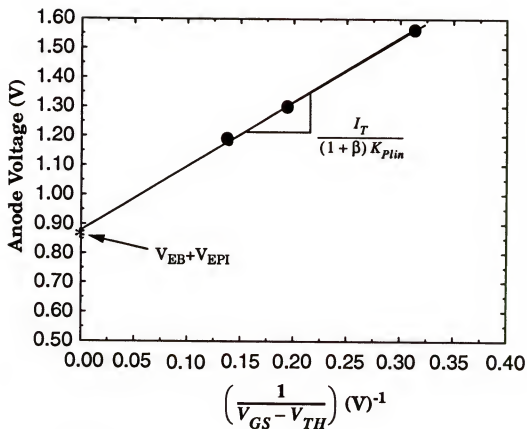


Fig.4.8 On-state voltage versus $1/(V_{GS} - V_{TH})$ at a constant anode current over the range of gate voltages. The slope is used to extract K_{plin} .

from this extraction step using the values of V_{TH} and β calculated from model parameters obtained in previous extraction steps.

4.5 Summary

The parameter extraction method for the LIGBT from simulation results have been developed and described in this chapter. It has been shown that unlike parameter extraction for microelectronic devices, the dynamic characteristics must be used because the characteristics of the internal LDMOST and lateral PNP bipolar transistor subcells are convoluted in the steady-state terminal characteristics of merged power devices such as the LIGBT. The dynamic waveforms contain many features that isolate different model parameters. The model parameters are obtained sequentially by selecting features of the device characteristics that isolate parameters, and using the parameters obtained from previous extraction steps to calculate the model parameters from the simulated characteristics.

CHAPTER 5 MODEL IMPLEMENTATION AND VERIFICATION

5.1 Introduction

A physics-based model for the low-gain, high-level injection lateral PNP BJT/LIGBT subcell combined with the LDMOST/LIGBT subcell model has been developed, resulting in a general purpose model for the LIGBT. The model has been verified for various external circuit conditions and for the full range of static and dynamic conditions in which the LIGBT is intended to be operated.

The purpose of this chapter is to describe the methodology for implementing the physical model of the LIGBT into the Saber circuit simulator [ANA92], and to provide a Saber LIGBT model that can be used for general purpose circuit simulations. Furthermore, the techniques demonstrated for modeling the LIGBT with the Saber circuit simulator are generally applicable to modeling other power semiconductor devices. Many diverse types of power devices are currently available with structures that differ significantly from one another, and different device model equations are generally required to describe each device type. This is in contrast to microelectronic devices for which many different devices can be described by using the appropriate model parameters in standard device model. This is the reason

that Saber became a very powerful modeling tool for power devices which have widely varying cell structures even within one category of device. The numerical machine in Saber is transparent to the user, and one template containing the model equations is the only implementation needed. This is not true of SPICE typically where, aside from the model subroutine, 17 other subroutines have to be modified. The techniques described in this chapter for implementing the LIGBT model equations into the Saber circuit simulator are essential for accurate simulation of power semiconductor devices. The model implementation into the Saber circuit simulator closely follows the treatment given by Hefner and Diebolt [HEF91], we will recapitulate the whole development for easy reference.

5.2 Implementing the LIGBT Model into Saber

To describe the behavior of a system such as an electrical network using the Saber circuit simulator, the interconnections of the different components of the system are described using a network list-ing (net-list). The net-list contains a statement for each component of the system that defines the name of the model template used to describe the component, the terminal connection points of the component, and the values of the model parameters that are to be changed from the default values of the generic model template. The model that describes each of the components of the system can be accessed from the Saber libraries of standard component models or from user-defined

Saber templates where the equations that describe the physical behavior of the device are implemented. The implementation of the LIGBT model equations into a Saber template is described in this section, and the net-lists describing the operation of the LIGBT within various test circuits are given in the next section.

5.2.1 Saber Template

Saber templates are written in the MAST modeling language which is similar to the C programming language with the addition of specially designed modeling constructs which facilitate the implementation of Kirchhoff's law and aid convergence. Both user-defined models and the standard Saber library models are implemented in Saber templates using the MAST modeling language. Electrical component models are implemented into templates by expressing the current through each element of the component in terms of the system variables of the component: system variables for electrical component models consist of terminal node voltages, internal node voltages, and explicitly defined system variables. The simulator solves for the system variables of the entire network such that the net current into each node of the system sums to zero. (i.e., Kirchhoff's current law is satisfied), and such that the equations defining the explicitly defined system variables for each components are satisfied.

A skeleton template of the Saber LIGBT model is shown in Fig.5.1 where each section performs the following functions: The LIGBT header defines the A (anode), G (gate), and K (cathode) termi-

```

Template LIGBT A G K = model
electrical A,G,K # node type
struct{
  number
    # user input parameters, and the parameter values can
    # be changed in the Saber net-list.
# local declaration
}
{
  parameters {
    # parameters calculated prior to simulation.
  }
  values {
    # values calculated as a function of system variables.
  }
  control {
    # simulator dependent control statements.
  }
  equations {
    # equations for system variables.
  }
}

```

Fig.5.1 Skeleton template of Saber LIGBT model.

nal connection points. The number section defines the user input parameters, and the parameters can be changed in the Saber net-list. The local declarations define constants, designate internal nodes, and explicitly define the additional system variables needed to describe the state of the device (usually define the sample point of the variables at each interval). The parameters section is used to calculate quantities that only need to be calculated once at the beginning of the simulation. Quantities that are functions of the system variables (explained in Chapter 3) are implemented in the value section. The control section contains information about nonlinear model relationships and commands to aid convergence. Finally, the equations section describes how the quantities calculated in the values section are assembled to solve for system variables.

5.2.2 LIGBT Model Formulation

To implement the LIGBT model presented in Chapter 3 into the Saber circuit simulator, the model is formulated such that the current between each of the terminal nodes are expressed in terms of the nonlinear functions of the system variables and in terms of the time rate of change of these functions of the system variables. The expressions in Chapter 3 are implemented in the values section of the Saber template, and these values are used in the equations section of the template to describe the interconnection of the components of current through each element in Fig. 5.1. Figure 5.2 shows a detailed LIGBT equivalent circuit superimposed on a schematic of the structure of one

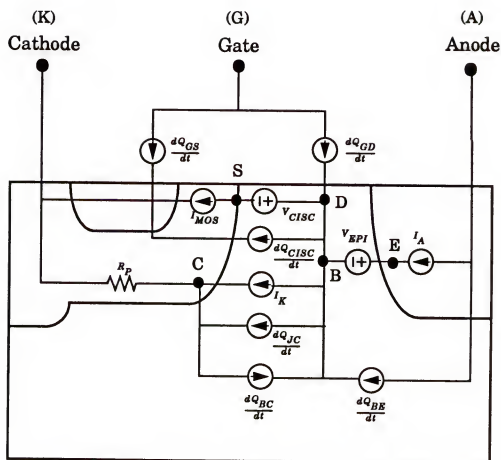


Fig.5.2 Network representation of the SABER LIGHT model superimposed on schematic of structure.

of the many thousand cells of an n-channel LIGBT. The elements of the circuit of Fig. 5.1 represent the nonlinear physical phenomena associated with each region of the device structure. The basic LIGBT equivalent circuit of a lateral PNP BJT/LIGBT subcell is supplied base current by the LDMOST/LIGBT subcell. The other components connected between the emitter (E), base (B), and collector (C) node are associated with the lateral PNP BJT/LIGBT subcell, and those connected between gate (G), source (S), and drain (D) nodes are associated with the LDMOST/LIGBT subcell.

The components associated with steady-state characteristics of the LIGBT model are I_A , V_{EPI} , V_{CISC} , I_{MOS} , and I_K . The potential drops due to drift and diffusion are distributed throughout the base region, and the drift terms of the ambipolar transport equations are coupled [HEF88]. Thus, both the base and collector components of current contribute to the resistive potential drop V_{EPI} . V_{CISC} represents the potential drop across current-induced-space-charge region. The current components of I_A , I_K , and I_{MOS} represent anode, collector, and LDMOST currents, respectively.

The components associated with dynamic characteristics can be divided into two groups: 1) the lateral PNP BJT/LIGBT subcell components, and 2) LDMOST/LIGBT subcell components. The components, $\frac{dQ_{BE}}{dt}$, $\frac{dQ_{BC}}{dt}$, and $\frac{dQ_{JC}}{dt}$ are associated with the lateral PNP BJT/LIGBT subcell, and the components, $\frac{dQ_{GD}}{dt}$, $\frac{dQ_{GS}}{dt}$, and $\frac{dQ_{CISC}}{dt}$ are associated with LDMOST/LIGBT subcell.

5.2.3 Saber Implementation

The expressions in Chapter 3 are implemented in the values section of the Saber LIGBT model template and are functions of the LIGBT system variables, the LIGBT device and model parameters (see Chapter 4), and the physical constants of silicon. The system variables for the LIGBT are the node voltages and the explicitly defined system variables: I_{EPI} and I_{CISC} . The first seven equations in the values section of LIGBT model template (see Appendix) evaluate the voltage differences. The notation $V(a)$ is the MAST syntax for the voltage at node a . The quantities evaluated in the values section are used in the equations shown in Fig. 5.3 to describe the currents through each of the elements of Fig.5.2, and to describe the expressions that define the explicitly defined system variables: I_{EPI} and I_{CISC} .

The first twelve statements in the equations section of Fig. 5.3 describe the currents between node pairs of Fig.5.2 in terms of the system variables. The notation $i(a \rightarrow b)$ indicates that a component of current, given by the expression on the right-hand side of $+=$, flows from node a to node b . The currents through the gate-source and drain-source capacitances are formulated as the time derivative of the charges Q_{GS} and Q_{DS} , where d_by_dt is the Saber time derivative operator, whereas the current components having no time derivatives such as I_{MOS} , I_{CISC} , I_A , and I_K are evaluated in the values section.

The last two statements in the equations section give the expressions that must be satisfied to determine each of the explicitly defined system variables. The notation of these statements indicates

```

equations {

    i(G->B)    +=d_by_dt(QGD)
    i(G->K)    +=d_by_dt(QGS)
    i(S->K)    +=IMOS
    i(B->S)    +=ICISC
    i(B->K)    +=d_by_dt(QCISC)
    i(A->E)    +=IA
    i(E->B)    +=IEPI
    i(A->B)    +=d_by_dt(QBE)
    i(C->B)    +=d_by_dt(QBC)
    i(B->C)    +=d_by_dt(QJK)
    i(B->C)    +=IK
    i(C->K)    +=VCK/RP

    IEPI:      v(E)-v(B)=VEPI
    ICISC:      v(B)-v(S)=VCISC

}

```

Fig.5.3 Equations section of Saber LIGBT model template.

that the system variable on the left-hand side of the colon (:) is to be solved for by the simulator such that the expression on the right-hand side of the colon is satisfied. That is, in addition to iterating the node voltages until Kirchhoff's current law is satisfied at each node. The simulator also iterates the explicitly defined system variables until each of the expressions on the right-hand sides of the colons are satisfied. System variables are introduced in this manner to describe components of current that cannot be expressed as explicit functions of the node voltages. System variables are also introduced to describe the time derivatives of capacitor voltages where the capacitance formula cannot be integrated to obtain an expression for the capacitor charge and hence the nonlinearities cannot be included within the argument of the Saber time derivative operator.

In the LIGBT template, the system variables I_{EPI} and I_{CISC} are introduced so that general expressions for the voltage sources can be readily implemented.

5.2.4 Techniques Used to Ensure Convergence

The nonlinear solution algorithm (iteration algorithm) of the Saber circuit simulator is unique in that all of the nonlinear expressions are evaluated in conjunction with an array of sample points for each of the independent variables. At the sample points, the nonlinear expressions are fully evaluated, while at intermediate values of the independent variables, the nonlinear functions are evaluated using multidimensional linear interpolations. One of the benefits of this

algorithm in implementing model equations is that the partial derivatives of the model equations with respect to the system variables are not required in order for the simulator to iterate the system variables in a manner that converges to the solution of the system of nonlinear equations. Thus, the model equations can be implemented directly in a straightforward manner as describe above.

The range and the density of the sample points can be tailored to the nature of the model nonlinearities by specifying the sample points arrays in the control section of the Saber template, which is shown in Fig.5.4. The simulator also provides the capability to control the maximum step size that a given variable can take between successive iterations (Newton step) in regions where the nonlinear model functions have discontinuous partial derivatives with respect to the system variables. This feature of the Saber simulator is beneficial in implementing models which use different expressions to describe different regions of operation. For example, I_{MOS} is described by a different expression for gate voltages above and below $V_{GS}=V_{TH}$. Newton steps for the independent variables can be introduced near the transition between different regions by specifying the Newton step arrays in the control section. The Newton steps tend to confine the iterations of the independent variables to the regions where they are introduced so that the variables do not overshoot the transition regions during iterations.

In implementing the LIGBT model into the Saber circuit simulator, the equations in Chapter 3 must be formulated such that they

```

struct sa_point {number bp,inc;}\

sva[*]=[(-100000,1.0),(0,0.001),(0.5,0.001),(1,0.1),(100000,0)],
sveb[*]=[(-100000,0.01),(0,0.001),(0.5,0.001),(50000.0,0)],
svch[*]=[(-50000,10),(0,0.1),(100000,0)],
svbk[*]=[(-50000,10),(0,.01),(1.0,1.0),(100000,0)],
nv[*]=[(4.5,0.01),(5.5,0)]

Control_section {

sample_points(VAK,sva)
sample_points(VEB,sveb)
sample_points(VBK,svbk)
sample_points(VCH,svch)
newton_step(VGS,nv)

}

```

Fig.5.4 Sample points and control section of Saber LIGBT model template.

are continuous and nonsingular in the range that the system variables may take during iterations. For example, the expression for X_D (base-collector depletion region width) is not valid for $V_{BK} < -0.6V$, and although no physical solutions exist in this range, it is necessary to provide a value in this range, because the base-collector voltage can enter the $V_{BK} < -0.6V$ range during the iterations necessary to find the physical solution. However, the values of the expressions in the range where there are no solutions are only important during iterations and have no effect on the results of the model. Therefore, the functions are given values outside the physical range of operation such that they are continuous at some point arbitrarily near the singularity, if it exists, and which are well defined beyond the range of physical operation.

5.3 Model Evaluation and Validation

The Saber LIGBT model described in the last section is evaluated using various test circuits to examine the device behavior for the range of static and dynamic conditions in which the device is intended to be operated. We have four general external circuits to test the Saber LIGBT model: 1) a series resistor-inductor load and resistive gate drive circuit, 2) flyback DC-to-DC converter, 3) soft-switching circuit with resonant inductor and capacitor, and 4) full bridge inverter employing four LIGBT's.

5.3.1 A Series Resistor-Inductor Load and Gate Drive Circuit

A series resistor-inductor load and resistive gate drive test circuit, shown in Fig. 5.5, is used to verify the transient model. The load circuit state equation for the circuit of Fig.5.5 is

$$\frac{dI_L}{dt} = \frac{1}{L} ((V_{AA} - R_L) \bullet (I_L - V_A)) \quad (5.1)$$

where $I_T = I_L$ for this circuit. The gate current for the circuit of Fig. 3.12 is given by

$$I_g = (V_{gg} - V_{gs}) / R_g \quad (5.2)$$

where the gate pulse generator voltage is given by

$$V_{gg} = \begin{bmatrix} 0 & \text{for } (t \leq t_{on}) \\ V_{gon}(t - t_{on}) / t_{rise} & \text{for } (t_{on} < t < t_{on} + t_{rise}) \\ V_{gon} & \text{for } (t_{on} + t_{rise} < t < t_{off}) \\ V_{gon}(t_{off} + t_{fall} - t) / t_{fall} & \text{for } (t_{off} < t < t_{off} + t_{fall}) \\ 0 & \text{for } (t \geq t_{off} + t_{fall}) \end{bmatrix} \quad (5.3)$$

The pulse generator rise and fall times are described by the second and fourth cases in Eqn.(5.3).

The corresponding net-list is shown in Fig. 5.6. The statements of Fig.5.6 describe each of the six circuit elements of Fig. 5.5, where the name of the given template that is used to model the component is given on the left-hand side of the period (.) and the name of

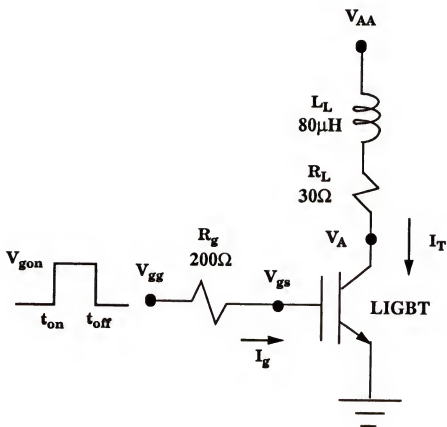


Fig.5.5 Circuit configuration of an LIGBT with a series resistor-inductor load and a resistive gate drive.

```

LIGBT.m1 a g 0= model=(dvw=0.1,tauh=2.6e-6)
r.r1 al a =30
l.l1 aa al =80u
r.rg gg g =0.2K
v.vg gg 0 = tran =(pulse=(0,15,0.5u,20n,20n,20u,40u))
v.va aa 0 = dc =300.0

```

Fig.5.6 Sabre net-list for a series resistor-inductor load and a resistive gate drive circuit.

specific component within the circuit is given on the right-hand side of the period. The resistor, inductor, and voltage supply templates are provided within the Saber template library, and the user-defined LIGBT template is described in Section 5.2. The remaining columns to the left of the equal sign define the connection points for each component, where the names of these connection points are defined in Fig. 5.5. The parameters of each component that are to be changed from the default values of the generic templates are given to the right of the equal signs.

Figure 5.7 shows the anode voltage waveform of Saber LIGBT model for different values of gate resistance. The value of the voltage overshoot at turn-off for a stiff gate drive ($R_g = 0.2 \text{ K}\Omega$) validates the implementation of base-collector depletion charge Q_{JC} which dominates the effective output capacitance. The turn-off delay time for the different values of gate resistances validates the implementation of the low-voltage gate-drain capacitance. The anode voltage overshoot for different gate resistances also validates the high-voltage gate-drain capacitance. Figure 5.8 shows the anode current waveform for three different values of gate resistance. The slowly decaying portion of the turn-off current waveforms validates the implementation of anode current I_A and non quasi-static charges Q_{BE} and Q_{BC} . The gate voltage and gate current waveforms for different values of gate resistance is shown in Fig. 5.9 and Fig. 5.10, respectively.

These characteristics are obtained using a Saber command file that allows you to perform a parameter sweep. In this parameter

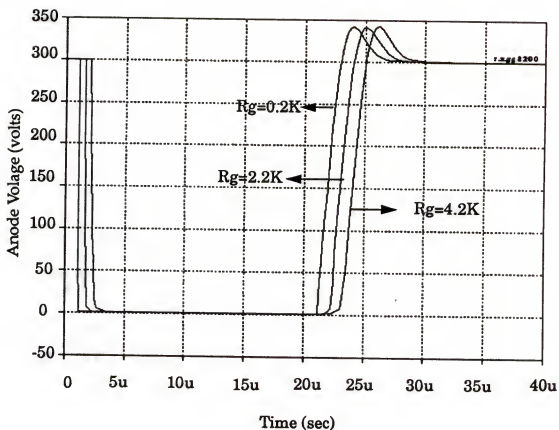


Fig.5.7 Saber simulated anode voltage waveforms for three different gate resistances ($R_g = 0.2K$, $2.2K$, and $4.2K$).

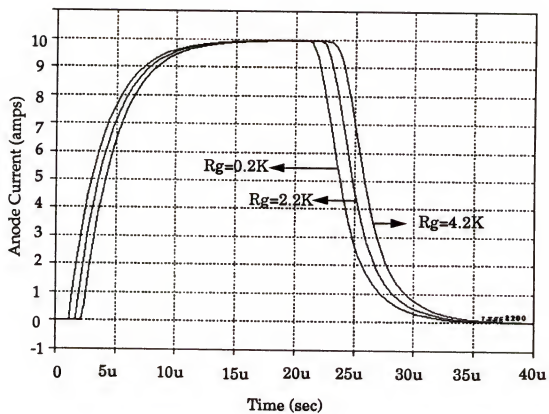


Fig.5.8 Saber simulated anode current waveforms for three different gate resistances ($R_g = 0.2K$, $2.2K$, and $4.2K$).

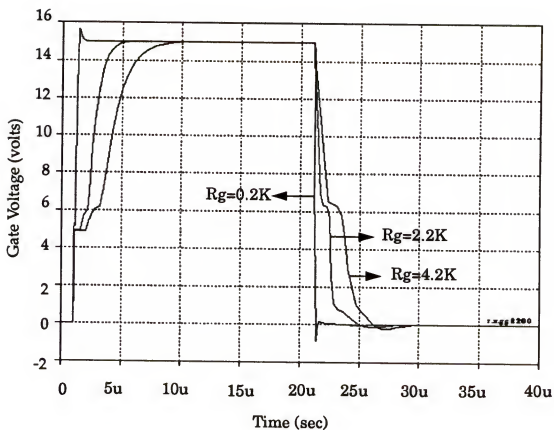


Fig.5.9 Saber simulated gate voltage waveforms for three different gate resistances ($R_g = 0.2K$, $2.2K$, and $4.2K$).

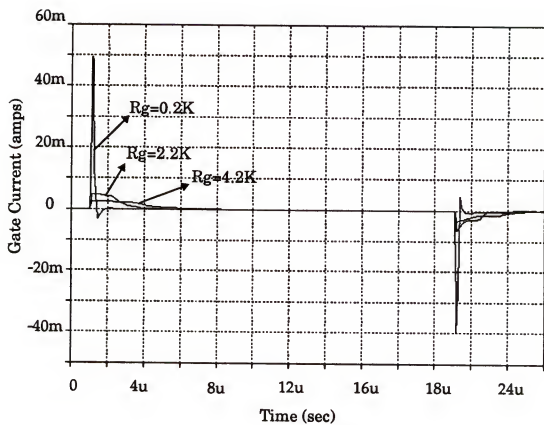


Fig.5.10 Saber simulated gate current waveforms for three different gate resistances ($R_g = 0.2K$, $2.2K$, and $4.2K$).

sweep, we use a range of gate resistance from 0.2K ohms to 4.2K ohms, with a step size of 2k ohms. This means that there is one execution of the transient analyses for each of the gate resistance values 0.2K, 2.2K, and 4.2K.

5.3.2 Flyback DC-to-DC Converter

The circuit of flyback DC-to-DC converter is shown in Fig. 5.11. When the LIGBT Q turns on, the diode D becomes reverse-biased and the energy flows into inductor core L. When Q turns off, the energy stored in the inductor core L causes the current flow through the diode in the counter-clockwise direction. In this flyback converter, the voltage-transfer ratio is

$$\frac{V_{out}}{V_{in}} = \frac{-D}{1-D} \quad (5.4)$$

where $D = t_{on}/T_s$ is the switch duty ratio. Eqn.(5.4) shows that the voltage transfer ratio in a flyback converter depends on switch duty ratio D. In this simulation, D is set to 1/2 so that the voltage transfer ratio becomes -1.

Figure 5.12 shows the anode-cathode voltage V_{AK} during the operation of the converter circuit. When the gate voltage is on, V_{AK} is almost zero (usually less than 2% or 3% of the peak voltage), and when the gate voltage is off, V_{AK} approaches nearly 240 volts. As a switching device, it is desirable that the LIGBT device has low on-state voltage and high off-state voltage across the anode and cathode terminals.

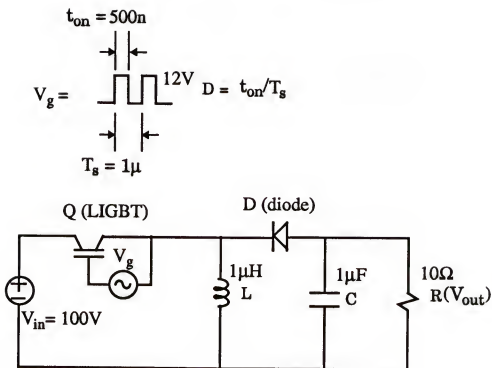


Fig.5.11 The flyback DC to DC converter circuit. The duty cycle is set $1/2$ to make output voltage of $-V_{in}$.

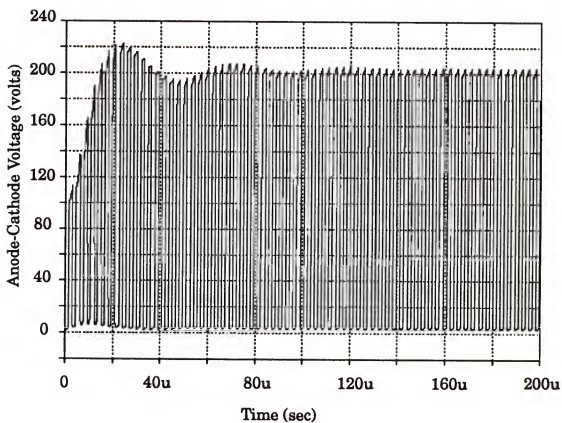


Fig.5.12 The anode-cathode voltage during on-off condition of the gate voltage.

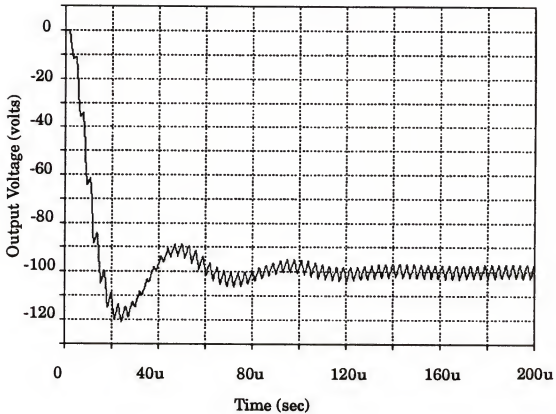


Fig.5.13 Output voltage of the flyback DC to DC converter. The ripple can be eliminated by increasing the value of capacitor.

The output voltage across the resistor R is shown in Fig.5.13. Two distinctive behavior is superimposed in this figure, the high frequency behavior by a switching circuit and low-frequency behavior of RLC circuit responding to step input. The high-frequency ripple can be reduced by increasing the value of capacitor C. The output voltage has a initial transient condition and settles down to steady-state of -100 volts.

5.3.3 Soft-Switching Circuit with LC

The application of soft-switching inverters is being actively considered by many manufacturers. Most available IGBT's are presently designed for hard-switching applications, very little data is available in the literature on device behavior under soft-switching conditions. The Saber simulation of soft-switching device has been motivated by a need to understand IGBT operation under the different operating conditions.

Figure 5.14 shows the soft-switching circuit with resonant inductor and capacitor. The test soft-switching circuit used for Saber simulation has been obtained from Kurnia et al. [KUR92]. The initial DC condition of the soft-switching circuit is $V_{out} = V_{in}$. When gate voltage is applied with a step function, the LIGBT turns on and the current in inductor increases. Therefore, there is no current charging the capacitor C. The inductor current is sinusoidal, but it is distorted when the LIGBT switching device is on. When the LIGBT turns off, the current in inductor flows to charge the capacitor, and the current

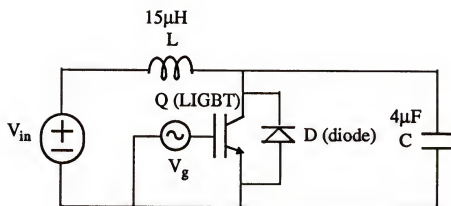


Fig.5.14 Soft-switching circuit with resonant inductor and capacitor.

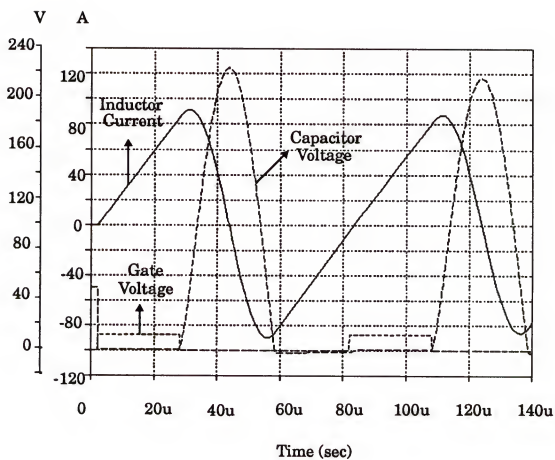


Fig.5.15 Saber simulated transient capacitor voltage, gate voltage, and inductor current waveforms.

waveform becomes sinusoidal. When the voltage across the capacitor becomes higher than the input voltage, the inductor current flows in the reverse way, which means the inductor current becomes more and more negative. Eventually, the capacitor voltage becomes zero as the inductor discharges the capacitor. Even when the voltage across the capacitor becomes zero, the inductor current still continues to flow its reverse way. This makes the parasitic diode D turn on, and the diode on-state voltage drop causes the negative voltage in capacitor.

When the inductor current approaches zero, the gate pulse generator is set to turn the gate voltage on to avoid hard-switching, and the whole cycle repeats.

5.3.4 Full-Bridge Inverter

A full-bridge inverter employing four IGBTs is shown in Fig.5.16. This inverter consists of two one-leg inverters and is preferred over other arrangements in higher power ratings. With the same DC input voltage, the maximum output voltage of the full bridge inverter is twice that of the half bridge inverter. This implies that for the same power, the output current and the switch currents are one-half of those for a half bridge inverter.

In Fig.5.16, the diagonally opposite switches (IGBT.M1, IGBT.M4) and (IGBT.M2, IGBT.M3) from the two legs are switched as pairs 1 and 2, respectively. Figure 5.17 shows the turn-off transient waveforms of the current and voltage for the IGBT.M2. The current remains steady while the voltage across the IGBT.M2 switching

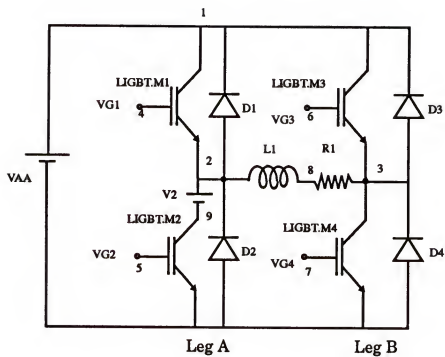


Fig.5.16 Schematic circuit of full bridge inverter employing four IGBTs.

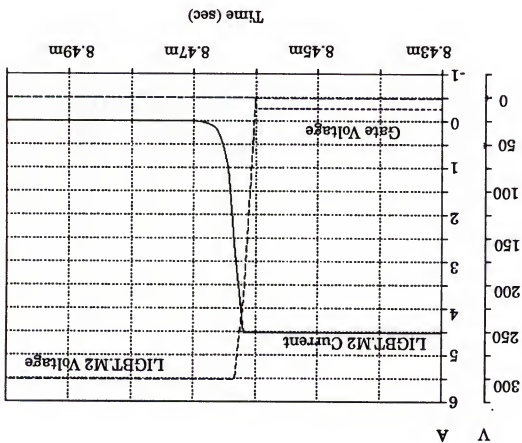


Fig.5.17 Turn-off transient waveforms for the LIGBT.M2.

device rises. Only after the voltage approaches its off-state value does current begin to fall. These are standard inductive switching waveforms as commonly seen in switch-mode power converters.

5.4 Summary

The physics-based model for the LIGBT has been implemented into the Saber circuit simulator. The techniques for implementing the LIGBT model equations into the Saber circuit simulator are essential for accurate simulation of other power semiconductor devices. The model has been verified for various external circuit conditions and for the full range of static and dynamic conditions in which the LIGBT is intended to be operated. The model has also been shown to be suitable for simulating the behavior of the LIGBT for general purpose external circuit conditions and designing the various power circuits.

CHAPTER 6

CONCLUSION AND SUGGESTIONS FOR FUTURE WORK

In this dissertation, an accurate charge-based model for the LIGBT has been developed and shown to be useful for optimal device/circuit CAD of IGBTs. Improvements on both subcell models of the LIGBT have resulted in accurate simulation results without parameter optimization. In the LDMOST/LIGBT subcell, we have modeled the current-induced-space-charge (CISC) region. The model of CISC region has been shown to be of great importance for improving the accuracy of the LDMOST model. In the lateral PNP BJT/LIGBT subcell, we have derived analytical geometrical factors based on the ambipolar transport equation. The lateral and quasi-vertical geometrical factors are needed in the two-dimensional anode current and thus in accurate modeling for steady-state and transient conditions for the LIGBT. The inclusion of the Kirk effect in our LIGBT model also improved the accuracy of steady-state characteristics especially for LIGBT with short distance between the anode and the cathode. The addition of Q_{BC} term in the LIGBT transient model also improved the accuracy of transient current and voltage waveforms of the LIGBT for general loading conditions.

The physical insights gained from two-dimensional numerical simulations and experimental measurements of test devices were used

to develop our charge-based LIGBT models. The generalized methodology for physics-based modeling of IGBTs based on the two-dimensional numerical device simulator PISCES can be useful to model the power devices with wide and lightly doped regions. The LIGBT model was implemented in the Saber circuit simulator, by which the model equations are solved semi-numerically within the nodal analysis framework of Saber. Saber is a very powerful modeling tool for power devices which have widely varying cell structures even within one category of device.

Our new model was verified by measurements of specially designed test devices, using simulations with model parameters evaluated from device structural information and static and dynamic measurements.

Parameter extraction method is developed. We, however, used PISCES simulation results due to unavailability of accurate measurement data. The model parameters are obtained sequentially by selecting features of the device characteristic that isolate parameters, and using the parameters obtained from previous extraction steps to calculate the model parameters from the next PISCES simulated characteristics.

The Saber LIGBT model has been used to simulate various external circuits and has been proven valid for simulating and designing the power electronic circuits.

Based on the research discussed herein, we suggest the following topics for future research consideration.

First, we recommend a study of regional models for IGBT. Each regional building block represents the physics-based model of each basic semiconductor region in the IGBT. Then IGBT model becomes a simple association of several of the basic semiconductor region models. These regional building blocks can be reused in modeling of other power devices with different structures such as PIN diode and power transistor.

Second, we recommend the refinements of LDMOST current model. The first-level of MOSFET current equation has been used in our LIGBT/LDMOST subcell model. More accurate description is needed in the model of the LDMOST/LIGBT subcell, for example to account for secondary and parasitic effects in the LDMOST channel region.

Third, we recommend the implementation of our LIGBT model directly into SPICE source code. The direct SPICE implementation requires the model routine to evaluate all the element values from the device terminal voltages.

Fourth, we recommend a study of the systematic parameter extraction with empiricism. The parameter extraction can be automated by linking ICCAP to Saber circuit simulator.

Fifth, we recommend the comparison between the measurement data of external test circuits shown in Chapter 5 and the Saber simulation results. This can be done either by obtaining actual multi-cell LIGBT devices or by modifying element values of the test circuits

based on the maximum current capacity of the single-cell LIGHT device.

APPENDIX A

LIGBT PISCES INPUT FILE

In this appendix, the PISCES input file for the LIGBT device provided by AT&T Bell Laboratories is listed. The PISCES input file details the geometry input and processing information of the AT&T LIGBT device.

PISCES INPUT FILE

TITLE LIGBT PISCES Simulation Input File

+ : Simulation Structure Generation

\$ Mesh Generation

```

MESH      rectangu x.min=0.0 x.max=55.0 y.min=0.0 y.max=20.0
+         smooth.k=1
X.MESH    x.left=0.0 x.right=8.5 h1=0.5 ratio=1.0
X.MESH    x.left=8.5 x.right=12.0 h1=0.25 ratio=1.0
X.MESH    x.left=11.0 x.right=39.25 h1=2.0 ratio=1.0
X.MESH    x.left=39.25 x.right=42.75 h1=0.25 ratio=1.0
X.MESH    x.left=42.75 x.right=55.0 h1=1.0 ratio=1.0
Y.MESH    node=1 location=-0.1
Y.MESH    node=4 location=0.0
Y.MESH    y.top=0.0 y.bottom=6.0 h1=0.25 ratio=1.0
Y.MESH    y.top=6.0 y.bottom=20.0 h1=2.0 ratio=1.0

```

\$ Mesh Elimination

```

ELIMINATE columns x.min=8.75 x.max=11.75 y.min=6.0
+         y.max=20.0
ELIMINATE columns x.min=8.75 x.max=11.75 y.min=6.0
+         y.max=20.0
ELIMINATE columns x.min=0.0 x.max=8.5 y.min=6.0 y.max=20.0
ELIMINATE columns x.min=39.25 x.max=42.75 y.min=6.0
+         y.max=20.0
ELIMINATE columns x.min=39.25 x.max=42.75 y.min=6.0
+         y.max=20.0

```

\$ Region Definition

```

REGION    number=1 oxide iy.min=1 iy.max=4
REGION    number=2 silicon x.min=0.0 x.max=55.0 iy.min=4
+         y.max=20.0

```

\$ Electrodes : #1=Anode #2=gate #3=Cathode

```

ELECTRODE number=1 x.min=45.0 x.max=55.0 iy.min=1
+         iy.max=4
ELECTRODE number=2 x.min=8.0 x.max=14.5 top
ELECTRODE number=3 x.min=0.0 x.max=7.0 iy.min=1 iy.max=4

```

\$ Substrate & Epi-layer Doping

```

PROFILE    n-type n.peak=2.0e14 x.peak=0.0 uniform y.peak=0.0
+         depth=20.0 outfile=hdop3.2

```

\$ P- Body Doping

```

PROFILE    p-type x.peak=0.0 x.right=8.0 n.peak=6.0e17

```

```

+           y.peak=0.4 y.juncti=4.0 xy.ratio=0.9

$   Deep P+ Doping
PROFILE   p-type x.peak=0.0 x.right=3.5 n.peak=1.0e19
+         y.peak=0.0 y.juncti=6.0 xy.ratio=0.8

$   N+ Source Doping
PROFILE   n-type x.peak=5.05 x.right=8.2 n.peak=5.0e19
+         y.peak=0.0 y.char=0.55 xy.ratio=0.65

$   P+ Anode Doping
PROFILE   p-type x.peak=45.0 x.right=55.0 n.peak=1.0e19
+         y.peak=0.0 y.junction=6.0 xy.ratio=0.8

$   Specification of Fixed Charge
INTERFACE qf=5.0e10

$   2-D Plot of Initial Grid
PLOT.2D   boun junc l.bound=2 l.elect=1 title="LIGBT Init"
+         x.min=0.0 x.max=16.0 y.min=-0.1 y.max=8.0
+         scale device=xterm pause

$   Regrid on Doping
REGRID    doping logarith smooth.k=1 ignore=1 cos angl=2.0
+         ratio=2.0 dopfile=hdop3.2
PLOT.2D   grid title="LIGBT Doping Regrid"
+         x.min=0.0 x.max=12.0 y.min=-0.1 y.max=8.0
+         device=xterm pause

$   Gate Contact Parameter
CONTACT   number=2 n.poly

$   Physical Model
MODELS    conmob fldmob srffmob2

$   Symbolic Factorization, Solve & Regrid on Potential
SYMBOLIC  carriers=0 gummel min.degr strip
METHOD    iccg damped
SOLVE     v1=0.0 v2=0.0 v3=0.0
REGRID    potentia smooth.k=1 ignore=1 cos angl=2.0
+         ratio=0.2 max.leve=1 dopfile=hdop3.2 outfile=pot3.mesh2
PLOT.2D   boun junc l.bound=2 l.elect=1 title="LIGBT Junction
+         Structure"
+         scale device=xterm

STOP

```

APPENDIX B

SABER LIGHT MODEL TEMPLATE

In this appendix, we overview the Saber implementation of the composite LIGHT model developed in chapter 3. All of the elements in the network representation of the LIGHT model shown in Fig.5.2 are implemented in Saber circuit simulator.

The source code of the LIGHT Saber template, which details the model routine for the LIGHT model shown in Fig.5.2, is listed on the following pages.

```
#####
#
# THIS WORK WAS SUPPORTED BY THE SRC UNDER CONTRACT
# 92-SJ-285. COPYRIGHT, 1993, UNIVERSITY OF FLORIDA.
# ALL RIGHTS RESERVED.
# FOR FURTHER INFORMATION, CONTACT :
# TAEHOON KIM AND D.E. BURK
# E-MAIL ADDRESS : THK@TCAD.EE.UFL.EDU
# OR DEB@TCAD.EE.UFL.EDU
# FEBRUARY, 1ST, 1995
#
#####
```

```
# light.sin  LIGHT.sin (Physical LIGHT)
# All parmeters are in M.K.S. Unit
# Temperature dependent parameters are installed
```

```
element template light A G K = model
```

```
#Header declaration
```

```
electrical A,G,K
struc {
```

```
#####
# User input paramter
#####
```

```
number T = 300.0
number PD = 5.0e22
number TAUH = 2.6e-6
number VTH = 4.8
number NEPI = 6.0e20
number AD = 1.0e25
number XJ = 6.4e-6
number WCH = 2.0e-6
number DVH = 16.0e-6
number PA = 10.0e-6
number ALJ = 8.0e-6
number WBV = 65.0e-6
number WB = 50.0e-6
number TOX = 1.0e-7
number DVW = 120.0e-6
number LSOL = 1.0e-6
```

```

number LDOL = 6.0e-6
number KPLIN = 0.96e-4
number KPSAT = 0.41e-4
} model = ()

```

```
{
```

```

#####
#                               Physical constants
#####

```

```

number QUE = 1.6022e-19 # Element charge
number PI = 3.14159
number BOLT= 1.38e-23   # Boltzman constant
number EPSO= 3.45e-1    # Oxide dielectric constant
number EPSS= 1.0e-10    # Silicon dielectric constant
number VS = 1.07e5      # Saturation velocity

```

```
#Values and variables declaration
```

```

number XKT,XP,REFT,PHIB,DNOMX,DNOMY,BR,COX,DN,DP,
      RTO,EG,NI,UN,UP,LA,TAUN,DA,JN0,RP,ARG,XLN,
      AREA,UNX,UPX,NEPIX,VT,WEPI,ADD,DNOMZ,UNZ,
      UNA,DNA,J1,MUC,TERM
val v   VAK,VEB,VBK,VCK,VEPIX,VEPIY,VEPIZ,VC,VEPI,
      VCISC,VCH,VGS,VGD,VBKL,VSBE
val i   IA,IMOS,IB,IK,IAL,IA0
val q   QGS,QGD,QBL,QBV,QJK,QCISC,QLAU,QLAD,QLA,
      QLBUL,QLBUR,QLBD,QLB,QBCL,QBEL,QBCV,
      QBEV,QBE,QBC
val meter XD,W,WX,XWM,WM
val nu    P0,XH,JAA,JAB,JA,JBX,JB,JMOS,JC,NCISC,XFAC,
      COFA,COFBA,QXT,AMOS,COFB,COFCA,COFCB,COFCBA,
      COFCD,COFCC,COFC,XINTA,XINTB,XINT,FGAL,CA,CU,
      CENY,MAXR,MINR,USEC,DSEC,PFGAV,SFGAV,FGAV,
      ACISC,CQL,PIC, QXTT,QBCLE,JK,XD1,BETA,LLA,
      ALPHA,JN,DPP,KPSAT,KPLIN
var i     iepi,icisc

```

```
#-----
```

```

group {VCISC,VEPI}      V
group {IA,IMOS,IB,IK}   I
group {QGS,QGD,QBC,QBE,QJK,QCISC} Q

```

```
#####
#                               internal nodes
#####
```

electrical E,B,S,C

```
#-----
```

```
      struc sa_point {number bp,inc;}\
```

```
sva[*]=[( -10000,1.0),(0,0.01),(0.5,0.01),(1,0.1),(10000,0)],
sveb[*]=[( -10000,0.01),(0,0.001),(0.5,0.001),(5000.0,0)],
svch[*]=[( -50000,10),(0,0.1),(100000,0)],
svbk[*]=[( -50000,10),(0,.01),(1.0,1.0),(100000,0)],
nv[*]=[(4.5,0.01),(5.5,0)]
```

```
#-----
```

```
PARAMETERS {
```

```
EG   = 1.16-(7.02e-4*model->T**2.)/(model->T+1108.0)
VT   = model->T*8.61e-5
XKT  = BOLTZ*model->T
ARG  = -EG/(XKT+XKT)+(1.11512784091/(BOLTZ*600.))
XP   = LIMEXP(QUE*ARG)
REFT = model->T/300.
NI   = 1.33e16*REFT**1.5*XP
PHIB = VT*LN(model->NEPI*5.0e21/NI**2.)
NEPIX = model->NEPI*1.0E-6
ADD  = model->AD*1.0E-6
DNOMX = 1.0+(NEPIX/(1.26e17*REFT**2.4))*0.88*REFT**-0.146
DNOMZ = 1.0+(ADD/(1.26e17*REFT**2.4))*0.88*REFT**-0.146
DNOMY = 1.0+(NEPIX/(2.35e17*REFT**2.4))*0.88*REFT**-0.146
UNX   = 88.0*REFT**-0.57+(7.4e8*model->T**-2.33)/DNOMX
UNZ   = 88.0*REFT**-0.57+(7.4e8*model->T**-2.33)/DNOMZ
UPX   = 54.3*REFT**-0.57+(1.36e8*model->T**-2.23)/DNOMY
UN    = UNX*1.0E-4
UNA   = UNZ*1.0E-4
UP    = UPX*1.0E-4
BR    = UN/UP
COX   = EPSO/model->TOX
DN     = XKT/QUE*UN
DNA    = XKT/QUE*UNA
DP     = XKT/QUE*UP
DA     = 2*DN*DP/(DN+DP)
LA     = SQRT(DA*model->TAUH)
```

```

RTO = model->DVW/model->WCH
MUC = UN/(1.+1.0E-6*(VGS-model->VTH))
TERM = MUC*RTO*COX
TAUN = 1./((1.+1.0e-17*ADD)/3.5e-5+8.3e-32*ADD**2.)
XLN = (DNA*TAUN)**0.5
JN0 = QUE*DNA**0.5/(TAUN**0.5*model->AD)
AREA = model->DVH*model->DVW
WEPI = model->DVH-model->XJ
Rp = 1./(model->PD*UP*QUE*model->DVW)
J1 = QUE*model->NEPI*VS
}

#-----

values {
#####
#           Node voltages
#####

VAK =v(A)-v(K)
VGS =v(G)-v(K)
VEB =v(A)-v(E)
VBK =v(B)-v(K)
VCH =v(S)-v(K)
VGD =v(G)-v(B)
VCK =v(C)-v(K)

#####
#           Calculating Anode current
#####

P0 =(NI**2./model->NEPI)*LIMEXP(VEB/VT)

IF(VBK<=0.0) {
XD=0
}
ELSE{
XD1=(2*EPSS*(PHIB+VBK)/(QUE*model->NEPI))**0.5
IF(JK<=0.0) {
XD=XD1
}
ELSE{
XD=XD1*(1.+JK/J1)**-0.5
}
}
W=model->WB-XD

```

```

JAA=JN0*(P0/NI)**2.
JAB=(QUE*DP*P0/LA)*1.0/TANH(W/LA)
JA=(BR+1)/BR*(JAB+JAA)
ALPHA=1.-(2.*DN/(DN+DP))*(1.-1./COSH(W/LA)))
BETA=ALPHA/(1-ALPHA)
DPP=DP

```

```

#####
#                               Calculating VEPI
#####

```

```

IF(IEPI<=0) {
  VEPI=0
}
ELSE {
  VSBE=VT*LN(model->NEPI**2./NI**2.)
  IF(VEB<=VSBE) {
    VEPI=0.0
  }
  ELSE {
    XH=model->NEPI*SINH(W/LA)/P0
    XWM=LA*LN(XH+(XH**2.+1.0)**0.5)
    VEPIY=(BR-1)/(BR+1)*VT*LN(P0/model->NEPI)
    WM=W-XWM
    IF(WM>=W) {
      VEPIZ=0.0
    }
    ELSE {
      VEPIX=JA*SINH(W/LA)*LA/(QUE*(UN+UP)*P0)
      VEPIX=JA*SINH(W/LA)*LA/(QUE*(UN+UP)*P0)
      VEPIY=(BR-1)/(BR+1)*VT*LN(P0/model->NEPI)
      VEPIW=(LN((TANH(W/(2.*LA))))
      VEPIV=VEPIW-LN((TANH((W-WM)/(2.*LA))))
      VEPIZ=VEPIX*VEPIW
    }
    VEPI=VEPIZ+VEPIY
  }
}

```

```

#####
#                               Calculating JB
#####

```

```

JBXX=(QUE*P0*LA/(model->TAUH*SINH(W/LA)))

```

```
JBX=JBXX*(COSH(W/LA)-1.0)
JB=JBX+JAA
JK=JA-JB
```

```
#####
#                               Calculating IMOS
#####
```

```
VC=VGS-model->VTH
IF(VC<=0.0) {
    IMOS=0.0
}
ELSE {
    IF(VCH<VC) {
        KPLINN=model->KPLIN*(model->DVW/120e-6)
        KPSATT=model->KPSAT*(model->DVW/120e-6)
        IMOS=KPLINN*(VGS-model->VTH-VCH/2.)*VCH
    }
    ELSE {
        IMOS=KPSATT*2.*(VGS-model->VTH)**2.
    }
}
```

```
#####
#                               Calculating VCISC
#####
```

```
IF(ICISC<=0) {
    VCISC=0
}
ELSE {
    JC=QUE*VS*model->NEPI
    IF(XD<=model->WCH) {
        WX=0.0
        VCISC=0.0
    }
    ELSE {
        WX=XD-model->WCH
        AMOS=0.5*PI*WX*model->DVW
        JMOS=IMOS/AMOS
        NCISC=JMOS/(QUE*VS)
        VCISC=0.5*QUE/EPSS*NCISC*WX**2.0
    }
}
```

}

```
#####
#                               Two dimensional analysis - Lateral side
#####
```

```
XFAC=LIMEXP(2*model->WB/LA)
COFA=(XFAC+1.0)/(XFAC-1.0)
COFBA=(XFAC+1.0)*XFAC*2.0*model->XJ/((XFAC-1.0)*LA)
COFBA=(2.0*model->XJ*XFAC/LA-COFBA)/(XFAC-1.0)
COFCA=1.0/12.0*XFAC*2.0*model->XJ*(12.0*model->XJ-LA)/LA**2
COFCBA=(12.0*model->XJ-LA)/((XFAC-1.0)*LA**2)
COFCB=1.0/12.0*(XFAC+1.0)*XFAC*2.0*model->XJ*COFCBA
COFCD=(2.0*model->WB/LA)**2
COFCC=LIMEXP(COFCD)*4.0*model->XJ**2/((XFAC-1.0)**2*LA**2)
COFC=(COFCA-COFCB+COFCC)/(XFAC-1.0)
PIC=PI/2.0
XINTA=COFA+COFB*PIC**2+COFC*PIC**4+24.0*COFC*(1.0+PIC)
XINTB=2.0*COFB*(1.0+PIC)+4.0*COFC*PIC**3+12.0*COFC*PIC**2
XINT=XINTA-XINTB
FGALL=model->DVW*XINT/(model->DVW+model->PA)
FGAL=FGALL*1./TANH(model->WB/LA)
```

```
#####
#                               Two dimensional analysis - Vertical side
#####
```

```
CA=model->WB/model->XJ
CU=WEPI/model->XJ
CENY=((model->WBV/2.0)**2-WEPI**2)/(2.0*WEPI)
MAXR=SQRT(CENY**2+(model->WBV/2.0)**2)
MINR=MAXR-WEPI
USEC=ACOS(model->XJ/MAXR)
DSEC=ACOS(model->XJ/MINR)
PFGAV=CA/2.0*(USEC+USEC**3/9.0+2.0*USEC**5/75.0)
SFGAV=CA/2.0*(DSEC+DSEC**3/9.0+2.0*DSEC**5/75.0)
FGAV=ABS(PFGAV-SFGAV)
```

```
#####
#                               Charging Terms
#####
```

```
QGD=(model->LDOL*model->DVW*COX)*VGD
QGS=(model->LSOL*model->DVW*COX)*VGS
ACISC=0.25*PI*WX**2.0
```

```

QCISC=0.0*QUE*ACISC*model->DVW*NCISC
CQL=LIMEXP(W/LA)
QLAU=0.5*CQL+1./(2.*CQL)-1
QLAD=0.5*CQL-1./(2.*CQL)
QLA=QLAU/QLAD
QLBUL=(0.5*(CQL-model->XJ)/LA)-(0.5*model->XJ/(LA*CQL))
QLBR=(CQL-1.)*(0.5*CQL*model->XJ/LA+0.5*model->XJ/(LA*CQL))
QLBUR=QLBR/(CQL+1.)
QLBD=0.5*CQL-(1./(2.*CQL))
QLB=(QLBUL-QLBUR)/QLBD
QXT=COSH(W/LA)-1.0
QXTT=0.5*(CQL**2+1.)/(CQL*SINH(W/LA))-(1./SINH(W/LA))
QBL=QUE*model->DVW*LA*P0*model->XJ*QXTT
QBCLE=0.5*(CQL**2-1.)*LA/(CQL*W*SINH(W/LA))-(1./SINH(W/LA))
QBCL=QUE*model->DVW*LA*P0*model->XJ*QBCLE
QBEL=QBL-QBCL
QBV=QBL*(FGAV/FGAL)
QBEV=QBV*(QBEL/QBL)
QBCV=QBV-QBEV
QBE=QBEL+QBEV
QBC=QBCL+QBCV
QJK=QUE*XD*AREA*model->NEPI

```

```

#####
#                               Calculating Anode current
#####

```

```

IA0=JA*(model->DVW+model->PA)*model->XJ
IA=IA0*(FGAL+FGAV)
IK=IA-JB*AREA
JK=JA-JB
IB=JB*AREA

```

```

}

```

```

#-----

```

```

Control_section {

```

```

sample_points(VAK,sva)
sample_points(VEB,sveb)
sample_points(VBK,svbk)
sample_points(VCH,svch)
newton_step(VGS,nv)

```

```

}

```

```

#-----
equations {

    i(G->B) +=d_by_dt(QGD)
    i(G->K) +=d_by_dt(QGS)
    i(S->K) +=IMOS
    i(B->S) +=ICISC
    i(B->K) +=d_by_dt(QCISC)
    i(A->E) +=IA
    i(E->B) +=IEPI
    i(A->B) +=d_by_dt(QBE)
    i(C->B) +=d_by_dt(QBC)
    i(B->C) +=d_by_dt(QJK)
    i(B->C) +=IK
    i(C->K) +=VCK/RP

    IEPI: v(E)-v(B)=VEPI
    ICISC: v(B)-v(S)=VCISC

}
}

```

REFERENCES

- [ADD84] Addler, M. S., Baliga, B. J., and Kokosa, R. A., "The evolution of power device technology," IEEE Trans. Electron Devices, vol. ED-31, pp. 1570, 1984.
- [ALL93] Allard, B., Morel, H., Lin, C. C., "Reusing basic semiconductor region models in power device bond graph definition," in Proc. of the Conf. EPE'93, Philadelphia, PA, pp. 34-39, 1993.
- [ANA90] Analogy, Inc., Guide to Writing Templates, Beaverton, OR, 1990.
- [APP79] Appels, J. A. and Vaes, H. M. J., "High-voltage thin layer devices (RESURF devices)," in IEEE IEDM Tech. Dig., pp. 238-241, 1979.
- [BEC85] Becke, H. W., "Approaches to isolation in high voltage integrated circuits," in IEEE IEDM Tech. Dig., pp. 724-729, 1985.
- [CHO92] Cho, H. J., "Charge-Based Model for the Lateral PNP Transistor," Ph.D. Dissertation, Univ. of Florida, Gainesville, 1992.
- [COU88] Coughran, W. M., Pinto, M. R., and Smith, R. K., "Computation of steady-state CMOS latch-up characteristics," IEEE Trans. Computer Aided Design, vol. 7, pp. 307-323, 1988.
- [DAR87] Darwish, M. and Board, K., "Lateral resurfed IGBT," Electronic Letters, vol. 20, pp. 5119-5220, 1987.
- [DIS93] Disney, D. R. and Plummer, J. D., "SOI LIGHT devices with a dual p-well implant for improved latching characteristics," Semiconductor Research Corporation (SRC) Publication C93135, 1993.

- [ERN91] Eranen, S., Gronlund, M. M., Blomberg, M., and Kiihamaki, J., "The Kirk effect in the LIGBT devices," IEEE Trans. Electron Devices, vol. 38, pp. 1919-1924, 1991.
- [FOS88] Fossum, J. G., Mcdonald, R. J., and Shibib, M. A., "Network representation of LIGBT structures for CAD of power integrated circuits," IEEE Trans. Electron Devices, vol. 35, pp. 507-515, 1988.
- [FOS86] Fossum, J. G. and Veeraraghvan, S., "Partitioned charge-based modeling of bipolar transistors for non-quasi-static simulation," IEEE Electron Device Lett., vol. EDL-7, pp. 652-654, 1986.
- [GHA77] Ghandi, S. K., Semiconductor Power Devices. Wiley, New York, 1977.
- [GOE94] Goebel, H., "Unified method for modeling semiconductor power devices," IEEE Trans. Electron Devices, vol. 9, pp. 497-505, 1994.
- [HEF93] Hefner, A. R. and Mantooth, H.A., "Component models for simulating power electronic circuits and system," tutorial course at IEEE PESC, Orlando, FL, 1993.
- [HEF91] Hefner, A. R. and Diebolt, D. M., "An experimentally verified IGBT model implemented in the Saber circuit simulator," in Proc. of the Conf. IEEE PESC, pp. 13-20, 1991.
- [HEF90] Hefner, A. R., "Analytical modeling of device-circuit interactions for the power insulated gate bipolar transistor (IGBT)," IEEE Trans. Industry Appl., vol. IA-26, pp. 995, 1990.
- [HEF88] Hefner, A. R. and Blackburn, D. L., "An analytical model for the steady-state and transient characteristics of the power insulated gate bipolar transistor," Solid-State Electronics, vol. 31, pp. 1513-1532, 1988.
- [KIM88] Kim, Y. S. and Fossum, J. G., "Static and dynamic latch-up in the LIGBT," IEEE Trans. Electron Devices, vol. 35, pp. 1977-1986, 1988.

- [KIR62] Kirk, C. T., "A theory of transistor cut-off frequency fall-off at high current densities," IRE Trans. Electron Devices, vol. ED-9, pp. 164-174, 1962.
- [KOS93] Koser, S. L., Shibib, M. A., Desto, J. C., "Combination field plate/field ring termination structures for integrated power devices," in 5th International Symp. on power semiconductor devices and IC's, Seattle, WA, pp. 182-187, 1993.
- [KUR92] Kurnia, A., Stielau, O. H., Venkataramanan, G., and Divan, D. M., "Loss mechanisms in IGBT's under zero voltage switching," in Proc. of the Conf. IEEE PESC, New York, NY, pp. 1011-1017, 1992.
- [LIN67] Lindmayer, J. and Schneider, W., "Theory of lateral transistors," Solid-state Electronics, vol. 10, pp. 225-234, 1967.
- [LU88] Lu, C. Y., Tsai, N. S., Dunn, C. N., Riffe, P. C., Shibib, M. A., Furnanage, R. A., and Goodwin, C. A., "An analog/digital BCDMOS technology with dielectric isolation-devices and processes," IEEE Trans. Electron Devices, vol. 35, pp. 230-239, 1988.
- [MA94] Ma, C. L., Lauritzen, P. O., and Budihardjo, I., "A systematic approach to modeling of power semiconductor devices based on charge control principles," in Proc. of the Conf. IEEE PESC, San Francisco, CA, pp. 31-37, 1994.
- [MAC87] McDonald, R. J., "A methodology for physical modeling of high voltage/power integrated circuit devices for circuit simulation," Ph.D. Dissertation, Univ. of Florida, Gainesville, 1987.
- [PAT86] Pattanayak, D. N., Robinson, A. L., Chow, T. P., Adler, M. S., Baliga, B. J., and Wildi, E. J., "n-channel lateral insulated gate transistors: Part I-steady-state characteristics," IEEE Trans. Electron Devices, vol. ED-33, pp. 1956-1963, 1986.
- [RUM85] Rummenik, V., "Power devices are in the chips," IEEE Spectrum, pp. 42, 1985.
- [SAB92] Analogy Inc., Saber Reference Manual, Beaverton, OR, 1992.

- [SIM85] Simpson, M. R., Gough, P. A., Hshieh, F. I., and Rummenik, V., "Analysis of the lateral insulated gate transistor," in IEEE IEDM Tech. Dig., pp. 740-743, 1985.
- [SHE93] Shen, Z and Chow, T.P., "Modeling and characterization of the insulated gate bipolar transistor (IGBT) for SPICE simulation," in 5th International Symp. on power semiconductor devices and IC's, Seattle, WA, pp. 165-170, 1993.
- [YIL84] Yilmaz, H., Van Dell, W. R., Owyang, K., and Zommer, N. D., "The insulated gate transistor (IGT)-a new power swtching device," in IEEE IEDM Tech. Dig., pp. 274, 1984.

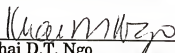
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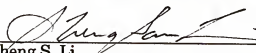
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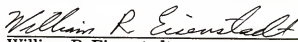
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